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new Sideband handbook

by don stoner, W6TNS

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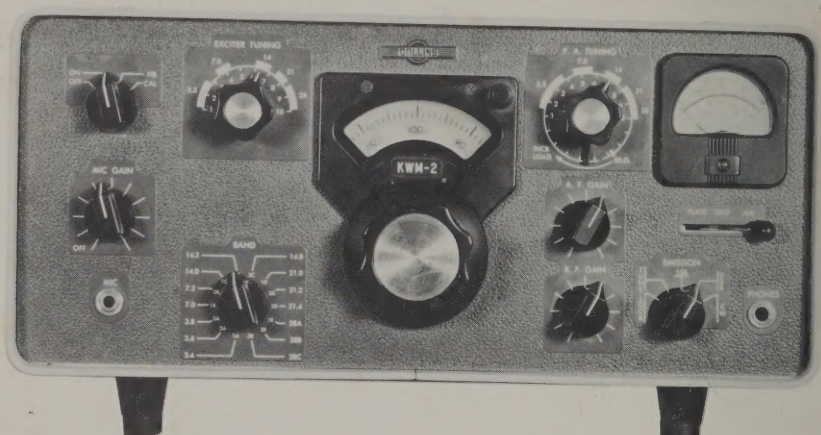

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by don stoner, W6TNS

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Introduction

ORDINARILY, this is the spot where the author tells you why you were wise to purchase his book. The background on the creation of the book, and credit lines usually finish out the page. I know that I seldom read the introduction because I am anxious to get into the "meat" of the book. And I suspect that the majority of the people that buy this book will do the same thing. So we shall skip the introduction as such.

But you, reading this now, are the exception. It is quite possible that you still operate AM phone, or possibly are just getting your feet wet with a new sideband rig. As long as you have taken the time to dwell on this page, I will use the space to introduce myself. Possibly this will start a new trend in introduction pages.

I should preface my remarks by advising you that I am a die-hard "sidewinder". Although a late comer to the ranks, it only took 10 minutes of SSB operation to convince me that this "new" form of transmission was "the berries". The first day on the air, I knocked off WAC two-way sideband! I might mention that I worked Greenland and several of the KC4 stations that day too, but it might sound like bragging and the ARRL does not count them anyway. The following weekend, the rest of the KC4 stations were polished off and I started on DXCC two-way sideband. I'm still working on that one though. It seems that there are a few countries that have not heard of sideband yet.

As I say, it took only 10 minutes but I really fell in love with sideband. For a guy that had never before worked into Europe or Africa, I was drunk with power. Maybe I actually was, for I gave away almost all of my AM gear!

If you have been snooping around in Single Sideband Handbooks and the various literature that appears from time-to-time, you have probably read about the 9 db advantage that SSB has over AM. Frankly, I have never measured it, in fact, I have never even tried. But I can tell you that when a sideband station switches over to AM, it is as if someone had pulled the chain and flushed the frequency. The db's went thataway. Hundreds of years ago a very ancient Greek philosopher once said the now famous words, "There a'int no free lunch", which freely translated means that you cannot get something for nothing. But the O.G.P. had never heard of single sideband. In this case the "free lunch" is in the form of free power, when side-

band is compared to ancient modulation (AM). For a given amount of usable radiated power, a single sideband transmitter will consume less power from the mains. Surely this indicates that the system is far more efficient. A little later in the book, I will try to explain why this is so. You can argue with my figures 'til you are "blue in the face" but you cannot argue with the fact that the sideband signals get through when AM will not. That is enough to sell me on it.

I understand there is a little difference of opinion about sideband. You often hear opinions brought out into the open at the club meetings. The longest and the loudest opinions usually come from the so-called "ole timer" who operates phone, AM style, that is. Strangely enough the good cw operators usually express a desire to try out sideband, where-as they would not even give AM a tumble. At club meetings one of these ole timers usually charge me like a beserk Brahma bull, with a not so cheerie greeting such as "So you operate sideband, huh"?. They usually have fire in their eyes and murder in their hearts. But I'm a skinny little guy and I wear glasses besides, so they don't scare me. After the blood recedes from their temples, a calm discussion usually brings out one or more of the following facts: 1) They have been "clobbered" on occasion by a "sidewinder". 2) Several hundred to several thousand dollars have been invested in AM gear and they are not about to give it away. 3) His communications receiver is lousy and he does not want to admit it. 4) The receiver is a good one, but he does not know how to tune it properly. And last, but most important 5) He does not understand sideband and therefore, has no recourse but to "knock" it. I cannot help you on numbers one, two, or three, but I sincerely hope that this book will clear up numbers four and five.

Speaking about the book (I knew that I could work around to it), every effort has been made to keep the theory as painless as possible. In some spots it might even be called interesting reading. I have attended many meetings on SSB and poured over the IRE Proceedings and the Single Sideband Issue for hours. These can be pretty frightening experiences for someone just getting their feet wet with a new sideband rig. With this thought in mind, the book was written.

Off on another tack, I would like to use a little of this space to thank my folks, particularly my father, K6BLC. Without his contribution to my knowledge of radio (and I might add, at times, superhuman patience and self control), this book would never have been written. Remember the time I cut the stamps off your QSL cards for my stamp collection, Dad? And I dare not forget my faithful XYL, Lucy. Without her tireless efforts in supplying pencils, coffee, cigarettes, and her constant nagging, nagging, this book would most surely have died along about Chapter II.

73, Don, W6TNS

1

Heath SB-10 SSB adapter.

What Is Sideband?

1.1 AMPLITUDE MODULATION AND THE CARRIER

QUITE a few years ago, when I was still building crystal sets, I heard or read the word "carrier". As I had often done before, I asked a question. "What's a carrier, Dad?" He replied "The carrier is the radio signal that you put your voice on. In other words, it carries your voice from one location to another via the ether, hence the name carrier." Obviously he was a phone man. Although not quite correct, it was a perfectly reasonable explanation, considering my age. I could visualize my words floating along on a mysterious magic carpet called a carrier. It was all quite clear to me. Sometime later, however, I happened to read about a radio transmitter that was used to transmit voice and Teletype simultaneously and did not transmit a carrier. At this point, I realized someone was "pulling my leg" and I started some concentrated reading on the subject of single sideband.

It is somewhat difficult to understand what a single sideband signal is composed of unless the complete theory of amplitude modulation is understood. For the purpose of an explanation, let us examine a typical transmitter. A simplified block

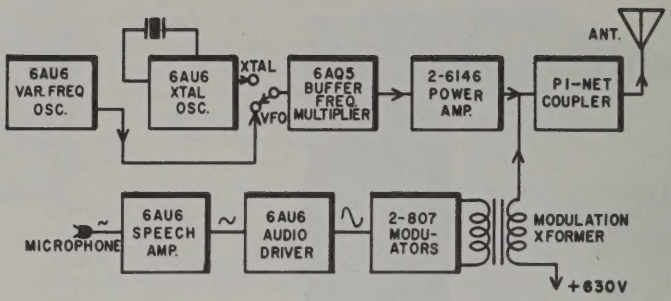


Fig. 1.1A—Block diagram of a typical amateur transmitter.

diagram is shown in figure 1.1A. Either the VFO or crystal oscillator generates a radio signal, or oscillation, which will eventually become the carrier. The rf from either of these oscillators is applied to a buffer/multiplier. This tube either isolates the oscillators from the power amplifier or multiplies the carrier frequency for the desired output. This steady continuous oscillation or carrier is amplified many times by the parallel 6146 power amplifiers. The signal going to the antenna from the pi-network consists of a very powerful oscillation on a single frequency. Naturally the antenna radiates this carrier to other antennas where it induces voltage. Keep in mind the fact that this transmitter radiates a single frequency, which we shall call the carrier. As an example, let's say the carrier is on 4 mc. If a telegraph key is used to start and stop the oscillations it is possible to add intelligence or information to the carrier by using a pre-arranged code consisting of a series of dots and dashes. I think we are all familiar with this system. In effect, the carrier (on 4 mc) is turned off and on and a receiver setup for cw reception and tuned to this frequency will reproduce the dots and dashes as an intermittent audio tone. In this manner intelligence can be transmitted from one location to another.

This is the basic form of radio transmission. Technically, the telegraph key can be called a modulator, for by definition it is used to add intelligence to the carrier. But what happens in our typical transmitter when it is modulated by voice? Referring back to the block diagram, you will note that the dc power to the final amplifier is connected in series with the modulation transformer. Audio voltage from the microphone is amplified by the speech amplifier and applied to the modulator, a pair of 807's operating in push-pull. The audio voltage representing our voice appears across the modulation transformer but amplified thousands of times by the speech amplifier, the driver and the modulator. The audio voltage also appears across the secondary of the modulation transformer, in series with the power supply voltage. Connected in this manner the audio will either aid or oppose the power supply voltage. Under conditions of 100% modulation the audio will double the voltage input to the final at the peak of the positive half cycle and the voltage will drop to zero at the peak of the negative half cycle. The drawing 1b in fig. 1.1B, illustrates 100% modulation. Compare this with the unmodulated carrier condition of a, fig. 1.1B. It can be seen that the instantaneous peak-to-peak voltage is twice that of the unmodulated carrier. This means that across a dummy load the peak voltage will be twice that of the unmodulated carrier. Since power equals the voltage squared divided by the resistance ($P=E^2/R$), where R is the dummy load resistance, the peak power at the extremes of the modulation cycle will be four times that of the unmodulated condition. All this is leading up to the following point. The final amplifier must be capable of delivering peaks of power four times greater than that of the carrier, or 400 percent. In an AM final amplifier, the voltage ratings

of the components must be much higher than they would be for cw, fm or sideband.

If the audio across the modulation transformer is insufficient, the percentage of modulation will be something less than 100% as shown in fig. c-1.1B. Conversely, if the audio is excessive the power input will try to increase beyond the capabilities of the transmitter. This means that during the negative half cycle the zero output condition will occur for a longer than normal period of time and peak clipping will occur. Also the positive half cycles may drive the final amplifier into saturation, thereby clipping the positive peaks. The peak clipping produces distortion and "splatter", hence the law requiring a maximum of 100% modulation. This condition, known as overmodulation, is shown in fig. d-1.1B. (Note the slight flattening of the positive peaks). It should be seen from the foregoing discussion that the power input of our typical transmitter is varied in direct proportion to our voice. Obviously, the power output (the radiated signal) will also vary with the voice, increasing or decreasing in amplitude as we talk. Basically, this is amplitude modulation.

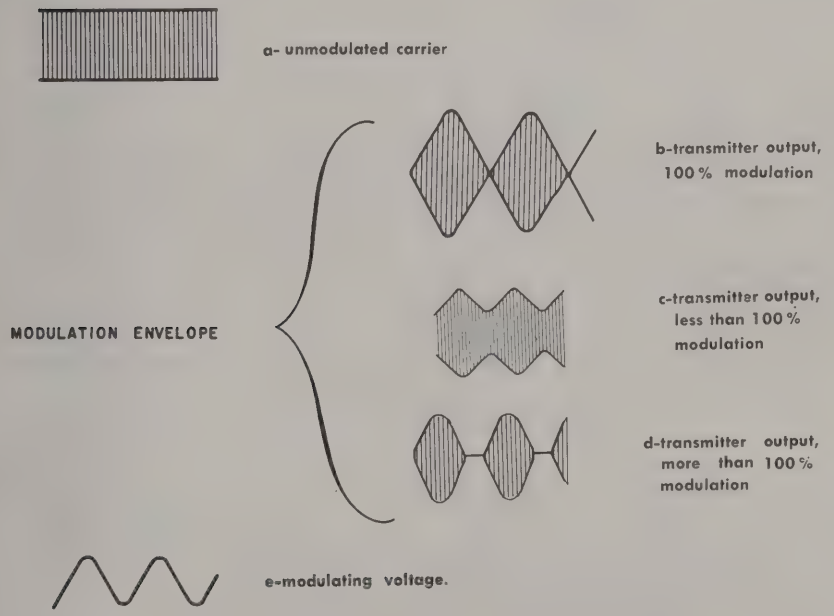
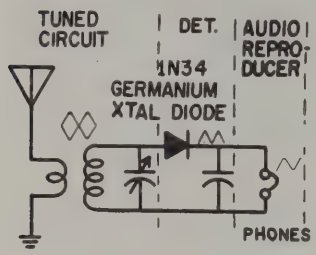


Fig. 1.1B—Typical transmitter waveforms.

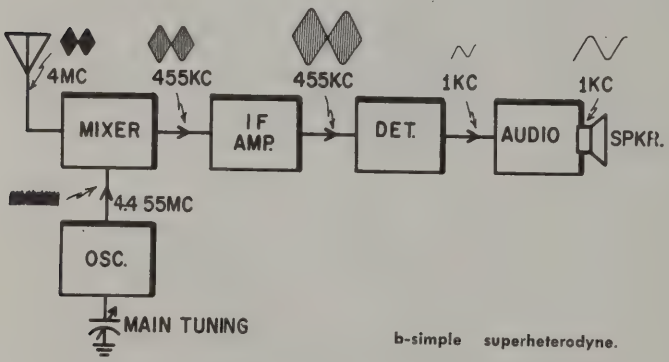
1.2 THE ROLE OF THE RECEIVER

The receiver is designed to convert the intelligence radiated by the transmitter to a form that can be heard and understood by the operator. In the process, the receiver performs the following functions: 1) it selects the desired station from the many signals that are propagated through the ether. 2) it removes the intel-

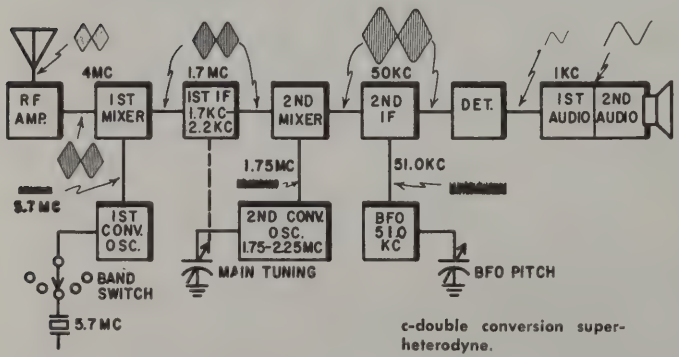
ligence from the signal, and 3) converts the electrical signal representing the intelligence to acoustical power that can be heard by our ears. The receiver may consist of a crystal detector (the simplest form of receiver), an inexpensive five tube superheterodyne, or a multistage double conversion superheterodyne. The crystal receiver and the inexpensive five tuber leave much to be desired in the selection of stations. The double conversion receiver usually represents the ultimate.



a-simple crystal detector.



b-simple superheterodyne.



c-double conversion superheterodyne.

Fig. 1.2—Receivers, from simple to deluxe.

Fig 1.2 shows these three forms of receivers in block and schematic diagram form. The diode detector consists of nothing more than a tuned circuit for separating stations, a detector for removing intelligence from the signal, and a set of headphones to convert the electrical energy to acoustical energy. The rf voltage induced in the antenna is coupled to the coil. The variable capacitor resonates the coil at the desired frequency. The rf voltage (similar to that shown in fig 1.1B b) is applied to the plate or anode of the germanium crystal diode. During the positive half cycle of rf energy the diode will conduct, since its plate is more positive than the cathode. This causes current to flow through the headphones. During the negative half cycle of rf the diode does not conduct and no current flows. In effect, we have chopped off the bottom half of the modulation envelope and utilized the top half. The capacitor across the headphones charges up to the peak of the rf cycle. The charge on the capacitor follows exactly the envelope as the modulation varies. This varying charge, or voltage across the capacitor, also appears across the headphones. The current flow through the headphones causes the diaphragm to vibrate at the same rate as the envelope, thereby reproducing the modulation. The diode detector meets the minimum requirements for a receiver, as outlined above, and that's just about all.

Before delving into the mysteries of the superheterodyne receiver, this is an excellent spot to explain some terms that will rear their ugly heads throughout the book. The words *heterodyne, mix, modulate,* and *convert* all mean the same thing. Whenever one term is used, it means just the same as the other three. This four headed process is the method whereby two signals of different frequencies are combined in some manner so as to produce two new signals of still different frequencies from the original two. The two new frequencies are referred to as the *sum* and *difference*. The medium in which the signals are combined must be non-linear such as your ears, a vacuum tube with no bias, or biased to cut-off and so forth. If it were perfectly linear, such as a true hi-fi amplifier would be, no mixing would occur. As a matter of fact, that is how hi-fi amplifiers are tested. Two signals on different frequencies are fed into the amplifier input and the amount of mixing is measured at the output. The amount of sum or difference signal is a direct indication of the amplifier linearity. A very similar process is used to test SSB linear amplifiers, too. The process of heterodyning, beating, mixing or converting will be found in almost any receiver or single sideband transmitter that you are likely to come across and for this reason this process should be thoroughly understood.

To examine what occurs in the simple superheterodyne, first let's put our typical transmitter on 4 mc and modulate it with a 1 kc audio oscillator. The diode detector described earlier has no amplification and the process of detection occurs at the transmitted frequency. It has been proven that more selectivity (ability to separate stations) will occur when amplification and detection is accomplished at a lower frequency. Any radio frequency may be converted to a lower frequency through the process known as *heterodyning*. Stated simply, heterodyning is the mixing of two frequencies to produce two new frequencies, the sum and difference. Have you ever noticed that when you whistle in the presence of another single audio frequency that a discordant sound is produced? This sound is the *beat frequency* sum and difference between the two sounds.

The superheterodyne operates on the same principle. As shown in fig 1.2b, the 4 mc signal, modulated by the 1 kc tone is applied to the mixer. Also applied to this tube is a steady signal from the local oscillator at a frequency of 4.455 mc. These two frequencies are combined in the mixer tube. In the plate circuit of the mixer four frequencies can be found. The original 4 mc signal, the 4.455 mc signals from the oscillator, an 8.455 mc sum signal, and a 455 kc difference signal. The lowest frequency is 455 kc and this is the frequency desired. The i-f signal is an exact replica of the incoming 4 mc signal except that the carrier frequency has

been changed from 4 mc to 455 kc. The modulation envelope, however, remains the same. This signal is amplified in the intermediate frequency stage and detected in the same manner as the diode detector mentioned earlier. The audio from the detector is sufficient to drive a pair of phones, but not a speaker. An audio amplifier is used to boost the level sufficiently to provide ample speaker volume. To change stations with this simple super-het the frequency is changed by varying the oscillator tank circuit. No matter what the frequency of the incoming signal, the difference frequency will always be 455 kc.

The more elaborate superheterodyne works in the same manner, but converts the signal twice rather than once as in the case of the simple receiver. It was stated earlier that the lower the frequency of amplification and detection, the greater the selectivity. But this also introduces other complications such as the reception of images, particularly when receiving above 10 mc. Although the incoming signal is below the oscillator, signals on the high side can also be heterodyned into the i-f amplifier. The unwanted signals are images. Even receivers with 455 kc i-f amplifiers are susceptible to image interference on the 10 meter band. We can "have our cake and eat it too" so to speak, by using double conversion. In this type of receiver a 1st i-f higher than 455 kc virtually eliminates the possibility of images. Then by converting down to a very low i-f, a high degree of selectivity is obtained. Fig. 1.2c illustrates a typical double conversion superheterodyne. The incoming 4 mc signal is amplified and presented to the mixer. The 4 mc signal is mixed with the crystal oscillator frequency of 5.7 mc and the difference of 1.7 mc is amplified in a tunable i-f stage. Stations are selected by tuning this stage. The 1.7 mc signal is again mixed with another oscillator operating on 1.75 mc. The difference signal is again utilized and amplified in a 50 kc stage (or stages). The detector and audio are similar to the simpler receiver. By using an i-f of 50 kc, it is possible to obtain a bandwidth in the order of 2,000 cycles without the use of crystal filters.

For code reception, the receiver is operated in a slightly different manner. The 50 kc signal no longer has variations in amplitude representing the intelligence, but rather, is a series of dots and dashes. These dots and dashes will not produce any audio from the detector. For cw reception, the receiver's beat frequency oscillator is turned on. This oscillator is displaced from the i-f slightly. In the case of fig 1.2c it operates on 51.0 kc. Each time the 50 kc incoming signal appears, the 51 kc oscillator beats with it, and produces a 1,000 cycle beat note that can be heard in the speaker. If the frequency of the bfo is changed to 50.5 kc the frequency of the beat note will become 500 cycles, and so on.

1.3 OTHER FORMS OF MODULATION

The typical transmitter described earlier incorporates a system of adding intelligence to the carrier, known as high level modulation. The power input to the final amplifier (and therefore, the signal amplitude) is varied by increasing or decreasing the plate voltage. Almost all Amateur class AM transmitters use this system.

It is also possible to control the power input with modulation by varying the gain of the tube. One method, popular a few years ago, was known as screen modulation and used either gate or clamp tubes to vary the screen voltage. It is also possible to vary the gain by changing the grid bias at an audio rate. This system is known as low level modulation, since it is in a low power level stage of the transmitter. Actually, any stage preceeding the power amplifier can be modulated so long as the stage is followed by amplifier(s) that are linear. A linear amplifier is one that faithfully reproduces any signal used to drive the stage. The tube is usually biased for Class AB to Class B which positions the operating range along the linear portion of the E_g-I_p curve. Because of the simplicity of high level plate

modulation, linear amplifiers were seldom used for Amateur applications until the advent of single sideband. It is easier to generate a single sideband signal at a low level stage and therefore, the linear amplifier is a must. Because this amplifier is so important to sideband applications, the entire contents of Chapter VI are devoted to the theory, construction and operation of linear amplifiers.

Frequency Modulation

This is the process of modulation whereby the carrier varies in frequency (rather than amplitude) at a rate determined by the modulation frequency. In effect, the carrier moves up from its normal resting frequency during one half cycle of modulation and below its normal resting frequency during the other half cycle. The distance (frequency-wise) the carrier moves is determined by the amplitude of the modulation voltage. The total deviation of a commercial FM station (88-108mc) is limited to $\pm$ 75 kc.

Pulse Modulation

The classic example of pulse modulation is, of course, a radar transmitter. The intelligence is the timing of a pulse chain. The amount of time it takes for the echo pulse to return from an object is directly related to the distance of the object from the transmitter/receiver.

Double Sideband

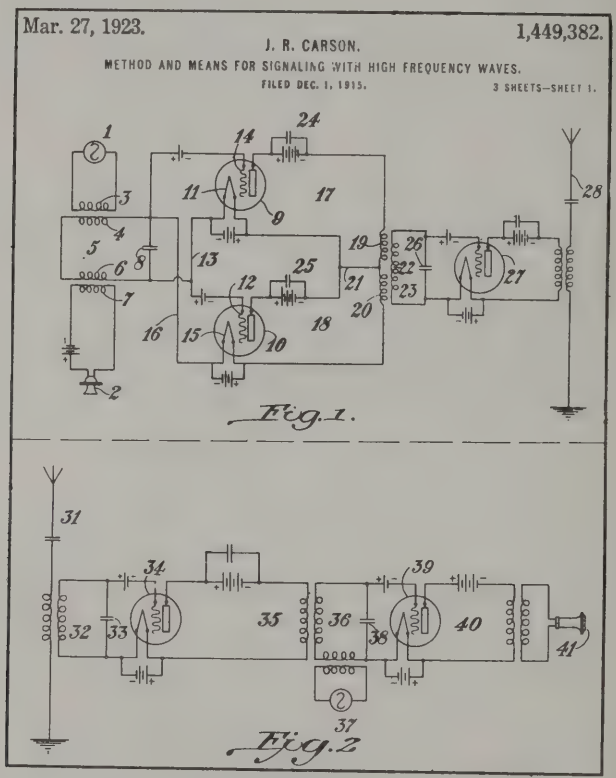
This is a form of transmission that has many of the advantages of single sideband, plus low cost and simplicity. It does, however, consume twice as much spectrum space as a single sideband signal. Double sideband is becoming increasingly popular with Amateurs and Chapter V describes the system and equipment in detail.

1.4 THE SIDEBAND CONCEPT OF PHONE TRANSMISSION

Contrary to popular opinion, sideband is not new, and was not invented in Cedar Rapids. In the year 1914 it was shown that something else occured in the modulated stage, beside amplitude variation. Whenever modulation was applied to the carrier frequency, the audio contained in the modulation heterodyned with the carrier to produce two new frequencies, the upper and lower sidebands. When the typical transmitter, described in 1.1, was modulated by a 1 kc tone the audio heterodyned with the 4 mc carrier to produce a lower sideband on 3999 kc, the carrier on 4 mc, and an upper sideband on 4001 kc. In the detector section of the receiver, the sidebands beat against the carrier and the sum and difference of 1 kc is amplified in the audio and heard in the speaker. The existence of the carrier and sidebands was proven during tests at NAA in 1915. Using extremely low frequency radio equipment, H. D. Arnold tuned the transmitting antenna to resonate at the lower sideband frequency, thereby attenuating the carrier and upper sideband.

Further tests were conducted by an employee of American Telephone and Telegraph Company, Mr. John R. Carson. Although the sideband concept was debated during this early period Carson reached two very important mathematical conclusions that formed the basis for single sideband as we know it today. One conclusion was that both of the sidebands contained identical information and in fact were mirror images of each other. His other conclusion was that of the energy radiated by the antenna, the greatest amount of power was contained in the carrier. The carrier contained absolutely no useful information other than as a reference for the sidebands! Carson also pointed out that if the carrier were removed the effective power output would be doubled with the same power input. For the same amount of power output, the power input could be cut in half thereby bringing about a healthy saving of power. To prove his point, Carson constructed a *balanced modulator* which cancelled out the carrier, but left the sidebands intact. He was

able to prove, to the satisfaction of AT&T that the carrier could be eliminated with no loss of intelligence. His work also included attempts at removal of the unwanted sideband. Using the original double sideband transmitter and filters between it and the antenna system, he was able to show that one of the sidebands could be eliminated without impairing the intelligibility of the signals.



A copy of the original patent obtained by John Carson in 1923.

A reference signal (to replace the carrier) was still needed, but Carson inserted it at the receiver detector. This locally generated carrier was on the same frequency as the carrier that had been suppressed in the transmitter but, of course, was many times weaker. The power and spectrum space saving afforded by his invention made possible the Trans-Atlantic submarine cable established in 1927. John R. Carson, considered the father of sideband, was granted patents on his system called "A Method and Means For Signaling with High Frequency Waves" in April 1923. Today sideband is used wherever a high efficiency, space saving, radio transmission is needed. The telephone company uses sideband to carry many conversations on each side of the suppressed carrier frequency. A typical system might include ten

simultaneous voice channels on each side of the carrier for a total of 20 circuits, all carrying different information. Some military gear places teletype circuits on one side of the suppressed carrier and voice channels on the other side. For maximum reliability, almost all overseas communication loops are transmitted by sideband. It is difficult for this writer to understand why Radio Amateurs have not pioneered this field. It is even more mysterious how some hams can argue the merits (and even the validity) of something the commercial services have taken for granted for more than 30 years.

1.5 SIDEBAND: PRO AND CON

This history of sideband should help to break ground for a somewhat more analytical discussion of sideband theory. Anyone who has operated sideband *knows* it is better than AM. But, unless the person has done a little thinking on the subject, he may not be sure *why* it is better.

Spectrum Requirements

Most amateur transmitters of the amplitude modulation variety are designed with restricted speech bandwidth. If the frequency response of the modulation system is "flat" within the audio range the intelligence contained in the sidebands would occupy more spectrum space than necessary. As an example, if your voice had 10 kc components, the sideband would extend 10 kc above and 10 kc below the carrier, or a total spectrum space of 20 kc. It has been established that only the voice frequencies between 300 and 3,000 cycles are required for intelligible communications. Not high fidelity by any means, but perfectly understandable. The telephone company restricts telephone conversations to these frequencies. No doubt you have noticed that a person's voice coming to you over a telephone does not sound at all like it would if they were in your presence. However, you have no trouble recognizing a particular voice and can mentally associate the person with that voice. Their speech is perfectly understandable. An amateur transmitter of good engineering design will contain a combination of high pass and low pass audio filters to restrict the modulator bandwidth below 300 cycles and above 3,000 cycles. Such filters need not be elaborate, and usually proper selection of resistance and capacity values in the audio section will suffice. Our typical AM transmitter would have sidebands above and below the carrier as far as 3,000 cycles from the carrier. Beyond 3,000 cycles the sidebands would rapidly diminish in strength. Such a transmitter would occupy about 6 kc of spectrum space.

A transmitter designed for single sideband operation would also contain a restricted audio bandwidth. Usually, more attention is given to the bandwidth of the speech stages to avoid defeating one of the main features of a single sideband transmitter, that is, the ultimate conservation of spectrum space. Many exciters contain resonant, or inductive filters to produce a sharp cut-off above 3,000 cycles. Since the single sideband transmitter radiates all its intelligence in one sideband the spectrum occupied starts about 300 cycles from the suppressed carrier and extends to about 3,000 cycles or a total spectrum space of 2,700 cycles. Naturally, sidebands do appear beyond these frequencies, but they are greatly attenuated with respect to the main "body" of the radiated signal. It is possible, by virtue of transmitting only one sideband, to place twice as many stations in a given band of frequencies.

Single sideband saves spectrum space in another way. This savings comes about very subtly. As mentioned earlier, the single sideband transmitter radiates only one sideband, no carrier and no unwanted sideband. By removing the sideband we have saved space, but by removing the carrier we have eliminated a terrific

source of interference. If you are active and on the air, you are undoubtedly very aware of what happens when another AM station transmits within a few kc's of the frequency you are listening on. A loud, and very annoying whistle is created by heterodyning or beating of the two carriers. When two sideband stations "get together" on adjacent frequencies, no nerve shattering heterodyne is created because no interference creating carrier is transmitted. For this reason, it is possible for single sideband stations to "snuggle up" to much less than 3 kc spacing. Certainly there is interference between the stations. So called "monkey chatter" can be heard on the station that you are listening to. But it is nothing like the grating sound of a heterodyne and far less annoying. Therefore, it is possible, with single sideband, to squeeze more than twice as many stations into a given band of frequencies. If you doubt the validity of this contention, examine the "sideband portion" of the 20 meter band, between 14.25 and 14.3 mc. If 3 kc were allotted to an information channel, then only 17 single sideband channels could occupy this portion of the 20 meter band. Actually, during peak activity hours, as many as 50 separate transmissions can be counted within that 50 kc band. If all these stations were to suddenly insert carrier, the result would be instant pandemonium! If 50 single sideband stations were placed on each one kc mark between 14.25 and 14.3, and each station had the same signal strength, it would be possible to copy every station with readability (and considerable monkey chatter).

Selective Fading

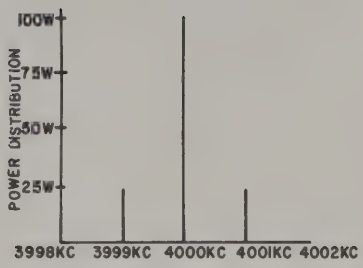
Although somewhat annoying, selective fading does not particularly plague ham operations. It can however, be very important in commercial services. This type of fading is caused when the carrier is received from two different ionospheric signal paths. If the two paths are a different length it will cause a phase difference between the two carriers. The time it takes for the signals to travel the two paths will determine the amount of phase shift between the two carriers. It happens quite often particularly in the evening and early morning hours when the ionosphere is unsettled. At times the phase of the two carriers will be 180° out and carrier cancellation will occur. The result is a horribly distorted sounding signal. In a few seconds the signal will return to normal. The ham operator might miss a few words but if it were a commercial or military transmission, those few words could be vital. One method of avoiding this type of phase distortion is dual diversity reception. Two receivers are tuned to the same frequency and are fed from separate antennas, with different polarization. Usually one receiver will have undistorted output while fading is observed from the other receiver. Parallel receiver outputs provide a continuous flow of undistorted intelligence.

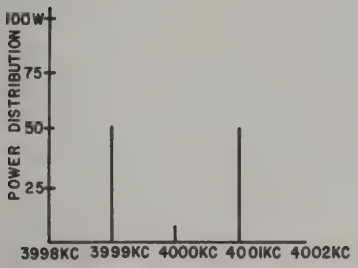
Another way of avoiding selective fading is to employ sideband. Since no carrier is transmitted there can be no out-of-phase carrier condition. The volume will appear to go up and down but the distortion products will not be present.

Free Power

The effective radiated power of a single sideband transmitter is many times that of a similar AM transmitter. To see how this is possible, let's refer back to our typical amateur transmitter. The transmitter is rated at 135 watts power *input*. With reasonable efficiency the power *output* is 100 watts. When the transmitter is modulated, the power varies at an audio rate. The average power is a combination of the dc and modulation power. With 100% sine wave modulation, the average power input would be 50% more than the dc power input or 202.5 watts input and 150 watts output. Referring to a of fig 1.5A you will see a spectrum drawing of what would be radiated by the transmitter if it were modulated by a 1,000 cycle tone. The power distribution is rather amazing. Of the 150 watts of power output, 100 watts is contained in the carrier and the carrier contains absolutely no intelligence but merely acts as a reference. The other 50 watts (supplied by the modulator) is

a-AM, 100% modulation;





b-DSB, 100% modulation,

c-SSB radiating the lower sideband.

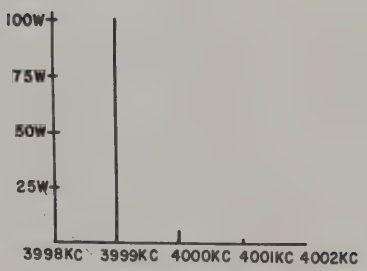


Fig. 1.5A—Power versus frequency distribution for several typical forms of transmission when modulated by a 1000 cycle tone;

contained in the sidebands (25 watts each) and they contain the same information, namely that a 1 kc tone is being transmitted. If someone told you that you were transmitting 25 watts of useful signal, with 135 watts input, you might think them looney. But this is exactly what happens. Don't blame your transmitter. Blame this outmoded and ancient method of transmission, it's the culprit!

To carry the discussion further, assume that the typical transmitter is modified in such a way that a signal containing the sidebands, and no carrier can be fed to the power amplifier. Remember that the power supply and final amplifier are capable of producing 100 watts output. No carrier is available for amplification, but the final is "straining at the bit" to amplify something and the available power is used to amplify the two sidebands. Each sideband will contain 50 watts now, as

shown in b of fig 1.5A. This is known as a double sideband signal and produces twice as much effective output as the AM signal with the same power input. Whenever power is doubled, it is said to have been increased by three db. This is the first "free" power increase over amplitude modulation on the way to that 9 db you have heard mentioned.

In the process of generating this AM signal, the plate voltage of the power amplifier is doubled at the peaks of the modulation waveform. Therefore, the peak voltage (carrier plus sidebands) will be twice the carrier and the peak power will be four times the carrier power. Whenever the voltage (E) is doubled, the power will increase by four times because $P=E^2/R$. This means that our typical transmitter, with an output power of 100 watts, and an average power (when modulated) of 150 watts must be able to handle a peak power of 400 watts.

Now let's go "whole hog" and feed the power amplifier with a signal containing no carrier and only one sideband. The power amplifier can still deliver 100 watts but the only signal available for amplification is the remaining sideband. The entire 100 watts is therefore contained in the sideband and the *effective* power output has again doubled for another 3 db, making a total, so far, of 6 db.

In our foregoing comparisons of the AM and SSB mode of transmitter operation we have neglected the slight difference in operating efficiencies of the class C (in the AM case) and the class AB_1 or B efficiency of SSB amplifiers. It is felt that this is only worthy of passing notice since the class C efficiency is about 75%, while the class AB_1 and class B amplifiers are only slightly less.

The last 3 db of the previously mentioned 9 db occurs in the receiver. It is a basic fact that the more selective the receiver, the better the signal-to-noise ratio will be. For AM reception the receiver would probably have a 6 kc bandwidth to pass the voice modulation frequencies without loss of intelligence. To copy a single sideband signal, the receiver bandwidth need only be half as wide, or 3 kc. In effect, this will improve the signal-to-noise ratio by an additional 3 db, making a total of 9 db over an amplitude modulation transmission system. There is considerable difference of opinion as to whether there is any signal-to-noise improvement in reducing the bandwidth in the presence of impulse type noise (such as automobile ignition, electric razors, brush type motors etc.) The nuisance value of this type of noise is strictly a subjective matter and has yet to be resolved to everyone's satisfaction. It might also be pointed out that a communications receiver can be adjusted to copy only one sideband and carrier of an AM station, too. So this last 3 db is subject to considerable debate. Rather than be accused of trying to claim 3 db that might not actually be merited, we will conclude (for now at any rate) that we can gain 3 db in the case of thermal noise and none in the case of impulse noise. How much of this total of 9 db is actually realized in operation? It depends entirely on band conditions as shown in fig 1.5B. Under ideal conditions, it makes no difference whether you are using AM, SSB or smoke signals. As the going gets a little rougher, SSB starts to take the lead and for marginal contacts during poor band conditions, SSB comes through like a knight in shining armor.

Single sideband also has several other "free power" advantages that may not have occured to you. For one thing, the modulators (power hogs usually) are eliminated. The reduction in power consumption will show up when it comes time to pay the electric bill. Another power saving feature of single sideband is the duty cycle that the tubes and power supplies are subjected to. The only time these components are "pushed" is during the peak portion of the positive cycles. In between the positive peaks, the power tubes and the power supply are allowed to "rest". Additionally, they rest between words, pauses in sentences and so on. Unlike the AM transmitter, which amplifies the carrier continuously, the single sideband transmitter "works" for short periods of time and rests for long periods of time. For this reason, the tubes will handle more effective power than would be possible in

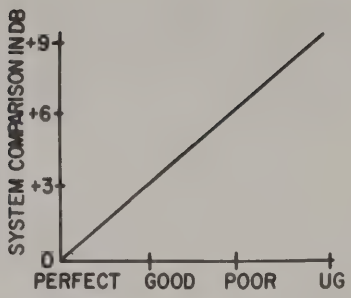


Fig. 1.5B—Comparison of SSB and AM with varying band conditions. The typical transmitter is used as 0 db reference.

AM service. Also, the high peak voltage created by the modulation is not present and the components in the power amplifier can be much smaller. Generally speaking, twice as much plate voltage can be applied to tubes in single sideband service as in AM service. Manufacturers usually avoid running tubes this close to their maximum ratings because in the hands of an inexperienced operator an improperly tuned transmitter can easily ruin the tubes.

The short duty cycle involved in transmitting single sideband also creates less exacting demands on the power supply. The heat created by the IR losses can be more easily dissipated during the "resting period" of the power supply. It is possible to obtain twice as much peak current from a power supply if the average current is held about 25% below the maximum rating of the supply. One may also correctly conclude that for a given power input a sideband power supply will be about half the size of its AM counterpart.

Television Interference

Contrary to what you may have heard, it is possible to have some rather severe TVI problems with sideband operation. Harmonic interference is definitely reduced because the stages used to boost the sideband signal to a usable level must be linear. Since the tube or tubes are not driven into the non-linear regions, harmonics of the fundamental frequency are not generated. An AM transmitter inherently generates TVI and the harmonics must be bottled up inside the rig.

The TVI problem with single sideband is created because of the terrific power that can be radiated. It has a tendency to *saturate* the front end of the receiver and to get into the 21 mc i-f amplifiers. A favorite cure for this type of TV set is to plug it into the 220 volt power mains. It fixes them every time! However, a more realistic solution is to install a high pass filter up close to the tuner and to shield the set if necessary. These techniques are the same as for AM, of course, and are covered more fully in other literature.

Scatter Contacts

This phenomenon is often experienced by Amateurs when operating sideband, and is worthy of mention under the advantages column. Although I am sure that it is possible to communicate via scatter with AM when enough power is employed, it was never experienced personally until operating sideband. Stations received via scatter have a very unusual "barrel sound", and can be easily recognized. The effect is most pronounced on the 20, 15 and 10 meter bands. Using scatter, it is possible to carry on a QSO with stations up to 1,000 miles away the same as you would on 80 or 40 meters. However, there is no severe fading and communication can be carried on any time of the day or night, even when the band is "dead". The signals are usually so weak that the "S" meter seldom reads up-scale except for a brief instant during a meteor ping. Scatter contacts can be carried on by pointing

both beam antennas at an imaginary spot several thousand miles away (back scatter) or by pointing both antennas toward each other (forward scatter). Signals are usually stronger on forward scatter, but in either case, they are "Q5 all the way".

Aurora Effects

Working into Europe on a propagation path over the North Pole is sometimes quite difficult. During periods of aurora disturbance a rapid flutter will be heard on signals arriving via this path. It appears to modulate the signal and at times it is so severe that communication with AM transmitters is impossible. The point where the earth's magnetic field lines converge at the poles tends to attract particles from outer space. The particles create an ionization curtain that is constantly varying and is influenced by the sun spot activity. Ever wonder why the KC4 (Antarctic) and KG1 (Greenland) stations use single sideband almost exclusively? It is because AM transmission will not provide the reliability necessary to get the traffic through, under conditions of aurora flutter. Again, this shows the advantage of SSB to get through when AM would not "cut the mustard".

Disadvantages of Sideband

The glory of double sideband and single sideband has been described in length. But what of the other side of the ledger? In a feeble attempt to appear completely unbiased, I will devote a minimum of space to the disadvantages.

The cost of getting on the air with a sideband signal is, of course, the biggest bug-a-boo. It is less expensive to build or buy a 100 watt AM rig, than a single sideband rig with the same average power input. Somewhere around 500 watts the cost is approximately the same and the cost curve crosses over to favor sideband at the kilowatt end of the curve. However, the prospective convert should keep in mind that adapters, such as the Heath SB-10, may be added to existing AM transmitters at a very low cost. Another easy way to get your feet wet is to purchase an inexpensive sideband exciter kit such as the Central Electronics 10B or the WRL DSB-100. A linear power amplifier can be easily constructed or the unit may be used "barefoot".

The extreme frequency stability requirements imposed by the sideband mode might be considered a disadvantage. Many receivers that perform beautifully when AM is being copied, leave the operator a shattered hulk after a few bouts with sideband reception. Manufacturers are becoming more aware of the stability shortcomings of their receivers and hams are now shopping for stability in addition to selectivity. Examined in this light, is it really a disadvantage? I prefer to think of

	100 Watt AM Transmitter	Same final amplifier used on SSB
Maximum Power on Voice Peaks	400 watt	400 watts
Average Power	100 watts (continuous)	Approx. 100-200 watts (only when talking)
Usable Sideband Power	25 watts per Sideband (2)	400 watts per Sideband (1)
System Bandwidth	6 Kc	3 Kc
Relative Receiver Output	1	4
Typical P.A. Plate Voltage	600 volts	1200 volts

A comparison between typical AM & SSB transmitters.

it in terms of bringing about an improvement in the quality of equipment and an advancement in the state of the art.

Another "tongue in cheek" disadvantage of sideband are the complications that go along with tuning the transmitter and/or receiver. Again, this is a new technique that must be learned and therefore is an advancement in the proficiency of the Amateur.

1.6 RECEIVING SIDEBAND

The method of intercepting and delivering sideband signals to the detector is the same as for AM. Frequency conversion is also identical although sideband reversal can occur in the conversion stage. For the purpose of the explanation, we shall assume that the sideband signal has progressed as far as the detector of the receiver. It was stated earlier that the carrier of an AM station heterodynes with the sidebands in the detector and we hear the beat notes at the recovered audio. But what about a single sideband signal? There is no carrier to beat the sideband with. The sideband signal will sound like the mating call of a pair of corduroy pants when heard on a receiver adjusted to receive AM.

To receive sideband properly, the carrier must be restored or "exalted" at the receiver. The carrier may be re-inserted anywhere in the receiver between the antenna and the audio detector. The common method is to insert the carrier just before the signal is demodulated by the detector. An older receiver can copy sideband if the beat frequency oscillator is turned on. The beat oscillator provides a tiny carrier that can be tuned across the i-f passband by varying the knob marked BFO Pitch. When this oscillator is adjusted to be on the same frequency as the carrier would be (if it were inserted) the original audio modulation can be recovered. If the oscillator is not on the same frequency as the suppressed carrier, voice components will be heterodyned to the wrong frequency. Thus, by varying the bfo, you can make the voice sound like "Donald Duck" or Vaughn Monroe with a head cold. To insert the correct frequency, the entire receiver must be both electrically and mechanically stable. Even a small frequency variation is noticeable as a variation in voice pitch. If the inserted carrier is off frequency by as much as 50 cycles, the voice components will shift. This means that the receiver should have an overall stability of 50 cycles. Rather a rough requirement, but entirely possible in an Amateur communications receiver of good engineering design.

It is possible to demodulate sideband signals by inserting the carrier at the antenna terminals. To do this, switch your transmitter to calibrate, or frequency spot and zero in on the sideband signal. It may be necessary to vary the coupling between the vfo and the receiver antenna terminals. The relative merit of these two systems will depend on whether your receiver or transmitter is the more stable and has the slower tuning rate.

In reading dissertations on how to tune sideband, you have no doubt been instructed to back off on the rf gain control. Why is this necessary? The carrier injection voltage from the bfo is frequently less than a volt. Usually only a few tenths of a volt, in fact. The sideband signal "coming down" the i-f strip will probably be several volts. When these two signals are mixed in the detector, overmodulation will occur. Remember the spectrum display of the typical AM signal? For 100% modulation the carrier contained twice as much power as the sidebands. When the AM transmitter was modulated more than 100%, the power in the sidebands increased while the power in the carrier decreased. When detecting a sideband signal, the bfo voltage will actually determine the modulation percentage. Unless the bfo signal (the carrier) has at least twice the voltage of the sideband, overmodulation will occur. Since few, if any, older receivers have provision for varying the bfo insertion, it is necessary to "back off" the rf gain control to reduce the

sideband voltage to less than half the bfo voltage. And it is necessary to constantly readjust the rf gain to compensate for signals of varying strength.

A device to "take you away from all this" is known as the product detector. It was invented by Murray G. Crosby, W2CSY, and is now standard equipment in sideband receivers. The product detector is self regulating in that the bfo injection voltage is always many times that of the incoming signal. This device is explained more fully, along with construction of a detector, in Chapter VII.

Almost any receiver is capable of demodulating sideband with a little practice, even the old one tube regenerative detector type. As a matter of fact, they work rather well! The ease with which you can tune sideband will depend on the design of the receiver, and can usually be gauged by the price.

1.7 APPARENT BROADNESS OF SIGNALS

Many hams are inclined to bellow loud and long about how broad single sideband signals are. Balder-dash and poppy-cock. "Hey fella, you're QRM'ing me and I'm operating 100 kc down the band" is occasionally heard. True, it is possible to create interference with a sideband rig if the operator is not entirely familiar with the controls. It is particularly true when an "eager beaver" is calling a rare one. He might be inclined to run the audio up too high and create distortion products. But to create spurious sidebands 100 kc away for any length of time is just about impossible. The sideband stations along side him would jump right down his throat after one or two transmissions.

What actually happens is this: The communications receiver is like a funnel, in that the i-f channel gets progressively more selective as the detector is approached. The more tuned circuits, the more selectivity, in other words. If the single sideband station is strong enough it is entirely possible to overload one of the stages just preceeding the detector. The rf is so high in amplitude that it drives the grid of the tube positive, causing it to draw grid current. The audible result is splatter on the station that you are trying to hear. The very simple solution to this dilemma is to back off on the rf gain control to reduce the i-f signal. The lost gain may be recovered simply by turning up the audio volume control. Actually this is good operating practice whether there is a sideband station operating near you, or not. An AM station can cause exactly the same type of interference. As a matter of fact, quite a bit of the blame for splatter can be attributed to poorly operated AM stations.

This first chapter of the book has attempted to give you a broad picture of sideband in as painless a manner as possible. To avoid repetition, several of the topics were skimmed over lightly, such as linear amplifiers, receiving and signal generation. These subjects are so diverse that complete chapters have been devoted to the theory and practical application.

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2

Balanced Modulators

JUST as the brush and paints are the basic tools of the artist, the balanced modulator is basic to a sideband transmitter. We learned in Chapter I that to heterodyne, mix, convert, or modulate, all meant the same thing. By the same token, a mixer, modulator, converter, or heterodyner (if I may be allowed to coin a new term) all accomplish the same purpose, that is, to combine two frequencies in order to produce two new frequencies. The new frequencies will consist of the mathematical sum and difference of the two original frequencies. The manner in which these two new signals are generated and utilized may vary widely. You are no doubt most familiar with the pentagrid tube used as the first mixer in most radio circuits. The balanced modulator can accomplish the same purpose as the pentagrid mixer. One might wonder, then, why complicate things by using balanced circuitry. The advantage of a balanced mixer over a single ended mixer is that the original modulating frequencies may be completely removed from the output circuit. As an example, if an rf carrier and audio modulation were applied to the single ended mixer the output signal would be identical to an AM transmitter and would consist of a carrier plus two sidebands. Since the balanced modulator only "passes" the sum and difference signals, the two sidebands will appear in the output. They are the sum and difference of the carrier and the modulation.

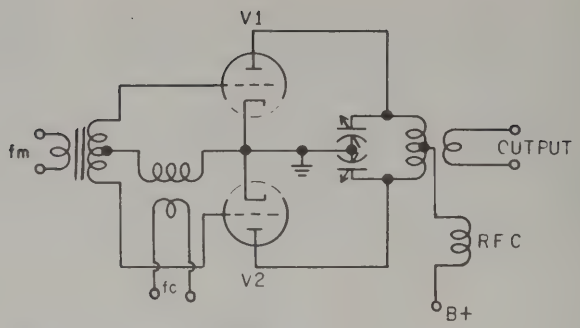


Fig. 2.1A—The basic push-pull balanced modulator circuit. A practical adaptation of this idea is shown in Fig. 2.1B.

2.1 PUSH-PULL BALANCED MODULATORS

The balanced modulator may be thought of as an electronic switch, wherein the carrier frequency is switched on and off at the rate of the modulating signal. Either vacuum tubes or metallic diodes may be used for such switching purposes. A simple balanced modulator circuit is shown in fig 2.1-A. At first glance this would appear to be a push-pull amplifier with some form of modulation applied to the center-tap of the grid coil. It may be considered as such except that the effects are slightly different than those normally encountered. The notations f_m and f_c denote the two frequencies that are fed into the balanced modulator stage—f_c being what we shall call the carrier frequency and f_m the modulating frequency.

For this example, assume that the modulating frequency, f_m, is the range of speech from 0 to 3000 cycles per second and the carrier frequency, f_c, is 1000 kilocycles. The push-pull tank circuit in the plate is tuned to a frequency of 1000 kilocycles. The following condition will then exist; With no speech modulation signal, f_m, present and the two halves of the push-pull circuit perfectly balanced, there will be no carrier output signal. It can be seen that the carrier voltage, f_c, is applied to the center-tap of the grid transformer and thus drives the two grids in parallel. If both grids swing positive simultaneously the individual plate currents of the two tubes will increase simultaneously. The effect in the plate tank circuit is that both pulses of plate current cancel each other out, making the net result in the output link winding zero. Thus any variation in grid signal voltage that is applied to both tube grids in parallel will cancel out in the plate circuit due to the push-pull circuit operation.

Consider now what the case would be if only the modulation frequency, f_m, were applied to the stage and the carrier frequency, f_c, were turned off for the moment. The speech signal is applied to the grids in push-pull connection and thus causes the tube plate currents to vary in a push-pull manner. Thus the plate current pulses of the triodes feed the tank circuit in the approved fashion of conventional push-pull amplifiers. There is one point, however, that changes the picture radically. The plate tank circuit is tuned to 1000 kilocycles and not to 0 — 3000 cycles. Therefore, the plate tank circuit does not present any load impedance to the speech frequencies and the pulses of plate current merely scurry on to the B-plus supply, causing no signal voltage to appear at the terminals of the output link coil. What has been accomplished thus far? By careful "balancing" we have eliminated the carrier signal in the output circuit, and by proper choice of tuned circuits used as

plate loads we have also prevented the modulating frequency from appearing at the output terminals. Thus, when either signal is present in the stage by itself, there is no output signal.

What happens when both signals are applied to the stage at the same time? The carrier, f_c, again feeds the grids in parallel, but the push-pull feed of the modulation causes one of the grids to swing less positive than the other. The amount is equal to the instantaneous peak voltage of the speech modulation voltage. The stage is no longer perfectly balanced and a signal voltage appears across the 1000 kilocycle tuned circuit. One might think that this signal voltage is made up of pulses of carrier signal and nothing else. However, if the signal at the output terminal is observed on a spectrum analyzer, it can be seen that it consists of signals grouped on either side of the carrier frequency and that no carrier signal is present. We have thus generated new signals whose frequencies are equal to the carrier frequency *plus and minus* the modulating frequency. In other words we have created two sidebands (one on each side of 1000 kc) and no carrier. These frequencies are equal to the carrier frequency plus the sum and difference, and may be termed a *double sideband suppressed carrier signal.* The balanced modulator is acting as a switch tube, turning the carrier, f_c, on and off at a rate determined by the modulating frequency, f_m.

The amplitude relation of the sidebands generated to the amplitudes of the two parent signals, f_c and f_m, is such that the output is proportional to the product of the two input signals. In somewhat simpler terms, if the modulating frequency signal doubled in amplitude, the output signal would be twice as large; however, if both the carrier and the modulating signal doubled in their individual amplitudes the output signal would be two-times-two or four times as great as the original. In the applications that we will deal with, the carrier signal will remain constant and the modulating signal will vary in amplitude causing the output signal to be linearly related to the modulating signal's amplitude.

There is one circuit condition that has not been mentioned up to this point. The tubes used in the balanced modulator must be operated on the non-linear portion of their grid voltage-plate current curve. This is quite contrary to the usual push-pull audio amplifier operation. Without this non-linear condition, the stage could not operate as described. Generally, in order to produce satisfactory results the carrier signal must be several times the amplitude of the modulating signal. This condition assures that the output signal will be linearly related to the modulation signal. The ratio of ten-to-one in carrier-to-signal amplitudes is generally accepted as the minimum value that will produce satisfactory results. Higher ratios are often desirable.

The circuit shown in fig. 2.1-A is the basic circuit described. This circuit does not have any provisions for balancing the "almost-always" existing tube and circuit unbalanced conditions. Fig. 2.1-B is the practical working circuit that should be used. The 1000-ohm cathode potentiometer provides the balancing adjustment necessary by adjusting the cathode biases on the two tubes. The modulating signal, f_m, is isolated from the carrier signal, f_c, by the pair of r-f chokes. This prevents the audio transformer from loading down the source of r-f signal.

2.2 THE MODIFIED PUSH-PULL CIRCUIT

Fig. 2.2-A shows an interesting modification that is possible on the push-pull arrangement. The carrier signal, f_c, is fed into the two cathodes in parallel feed causing cancellation in the push-pull connected plate circuit. The two triodes are thus operated in grounded-grid fashion for the carrier signal. The grid of the upper triode is at r-f ground potential because the reactance of the grid capacitor is low

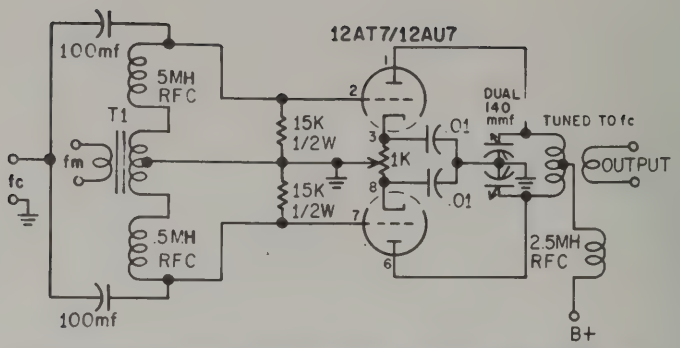


Fig. 2.1B—Typical practical push-pull balanced modulator. This circuit is quite popular and is found in SSB adapters such as the B&W 51SB and the Heath SB-10.

at the carrier frequency, f_c. The audio signal f_m, is not fed to the grids in push-pull as in fig. 2.1-A, but is fed to only one grid by an R/C coupled circuit. The upper triode acts as a cathode follower and the signal (f_m) appears across the common cathode resistor (3.3K) which varies the bias for the lower triode. In this manner the conduction of each triode is varied as f_m alternates. This circuit also eliminates the need for the interstage transformer of fig. 2.1-B.

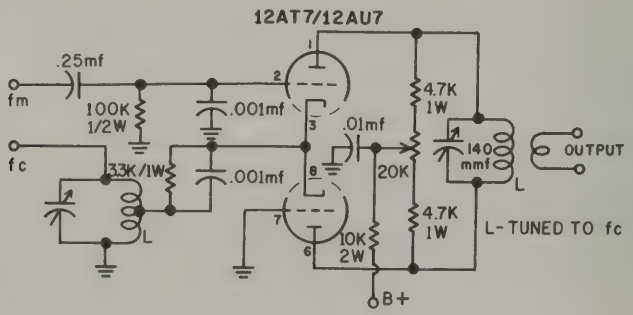


Fig. 2.2A—Modified push-pull balanced modulator.

2.3 THE PUSH-PUSH CONNECTED CIRCUIT

Fig. 2.3-A is very similar to the push-push frequency doubler that has been used for years. The plate tank circuit, however, is not tuned to twice the grid input frequency but is tuned to the fundamental of the grid r-f signal. It will be noted that the r-f carrier signal, f_c, is fed to the grids in push-pull while the modulating signal, f_m, is fed to the two grids in parallel through the grid coil center tap.

The fundamental frequency of f_c will cancel out in the tank circuit because each pulse from the parallel connected plates will oppose the "filled-in" half-cycle of the r-f wave form furnished by the fly-wheel effect of the tuned tank circuit. Thus, if the two tubes are perfectly balanced, no carrier signal will appear in the output circuit. It should be mentioned here that due to the push-push connection,

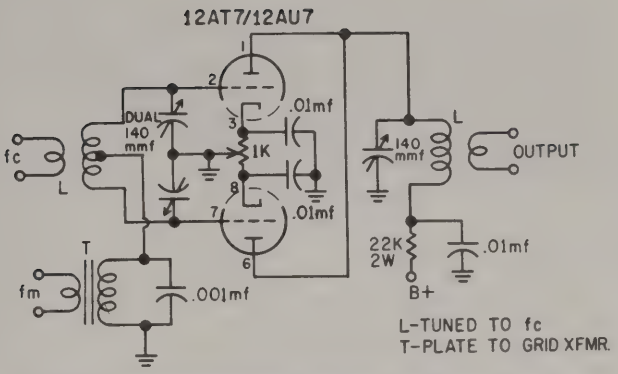


Fig. 2.3A—A push-push balanced modulator with both grid and plate circuits tuned to the same carrier frequency.

the even harmonics (i.e., the 2nd, 4th, 6th, etc.) will tend to be accentuated and their appearance in the output will depend on the tuned circuit "Q". (See the section on "Spurious Signal Generation").

Since modulating voltage is connected to the grids in parallel the current of the two parallel connected plates will respond directly. However, again as in fig. 2.1-A the plate tuned circuit will not respond to f_m so it will not appear in the output. When f_m and f_c are simultaneously applied to the stage the result is the same as for the first type described, that is, a double sideband suppressed carrier signal is generated in the plate circuit.

2.4 THE "UNBALANCED" BALANCED MODULATOR

The circuit shown in fig. 2.4-A is the so-called "unbalanced" balanced modulator. Actually the circuit takes advantage of the fact that vacuum tubes can shift phase by 180°. It is an extremely interesting and useful circuit in that it does not use a push-pull signal source or tank coil. The carrier signal, f_c, is applied to the grid of V1 at say zero degrees, as an example. This section of the 12AU7 serves

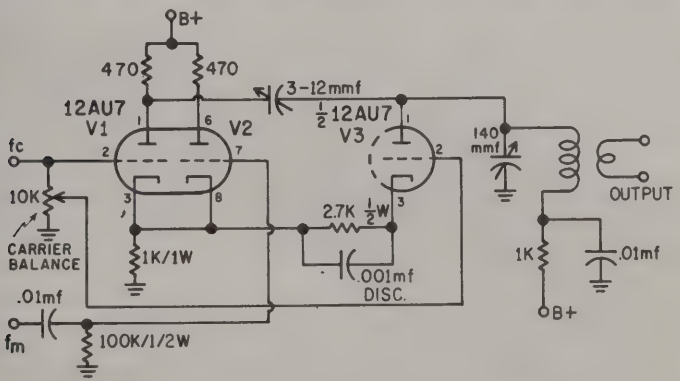


Fig. 2.4A—A clever balanced modulator employing no push-pull input signals or output circuits.

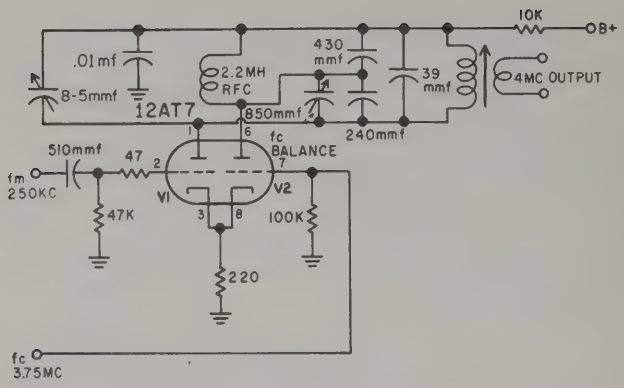


Fig. 2.4B—Another version of the balanced modulator that does not require push-pull circuits. This circuit is employed by Collins, in the KWS-1 transmitter, for frequency conversion.

as an r-f cathode follower. The carrier signal develops across the 1K cathode resistor (still at 0°), and is coupled to the cathode of V3. This stage operates similarly to a grounded grid amplifier, and the carrier appears across the output tank circuit "in phase" with the original f_c. You will note that the carrier is also applied to the carrier balance potentiometer and a small portion of this signal is applied to pin 2 of V3. This tube (V3) amplifies f_c and shifts the phase by 180°. By proper adjustment of the carrier balance control it is possible to obtain a null in the carrier signal. The 3-12 mmf. capacitor will correct for phase errors to produce near perfect carrier cancellation. Once this capacitor has been set, it need not be changed unless the tubes are replaced.

When modulation, f_m, is applied, the current flow in the audio cathode follower (V2) will vary in proportion to the audio. This in-turn changes the voltage across the 1K cathode resistor, varies the bias on V1, and unbalances the carrier. The output consists of the sum and difference signals (the sidebands), but no carrier.

Another form of balanced modulator that does not require push-pull circuitry is shown in fig. 2.4-B. This circuit is used extensively in the Collins KWS-1 transmitter for signal mixing. The ratio of the voltages on pins 2 and 7 is roughly 0.1 volts to 1.5 volts. The 3.75 mc signal is several times the 250 kc input. This necessitates a rather complex tuned plate circuit to reject the strong undesired signal in the output tank. The theoretical principle involved is selective feedback rejection. The 8-50, 240, and 430 mmf. capacitors form a divider for the feedback voltage. The variable capacitor in the feedback network is adjusted for a null of the particular frequency to be rejected.

It is worth noting that this circuit will also work very well if the 250 kc f_m signal is replaced with audio modulation. During positive cycles of modulation, current in V1 will increase while current in V2 will decrease, thereby unbalancing the system. A double sideband, suppressed carrier signal will appear in the output circuit.

2.5 MULTI-GRID TUBES IN BALANCED MODULATOR CIRCUITS

The circuits described so far have made use of triodes only. The multi-grid tubes may also be used and the isolation offered by the separated grids makes them attractive for some applications. Basically the operation is the same as for the types

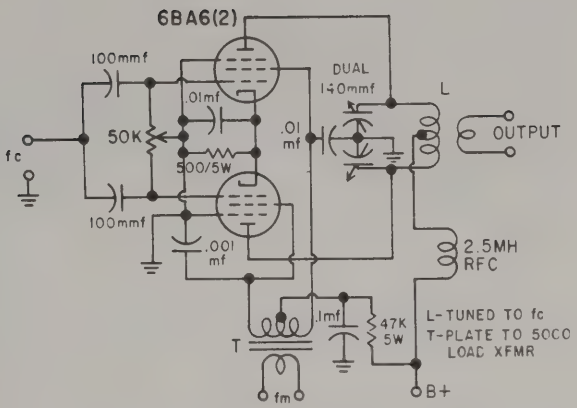


Fig. 2.5A—A balanced modulator using pentodes. The audio signal is applied to the screen grids to "switch" tubes.

already described. The r-f carrier is applied in parallel to the control grids in the case of the pentodes (fig. 2.5-A) and to the oscillator injection grids in the case of the pentagrid converter tubes (fig. 2.5-B). When the stage is balanced the carrier is non-existent in the plate circuit. The modulating voltage is applied to the screen grid of pentodes and to the signal grid of the pentagrid converter. Application of the modulation disturbs the balance of the system by making one plate current vary in a different manner than the other. The balancing potentiometer control operates by varying the proportion of the excitation voltage that is applied to each grid. Since both ends of the potentiometer are at the r-f carrier potential, it must be located near the grid terminals of the tube. This may prove a disadvantage in some instances. The cathode balancing circuit of fig. 2.1-B is always at zero r-f potential and may be located remotely.

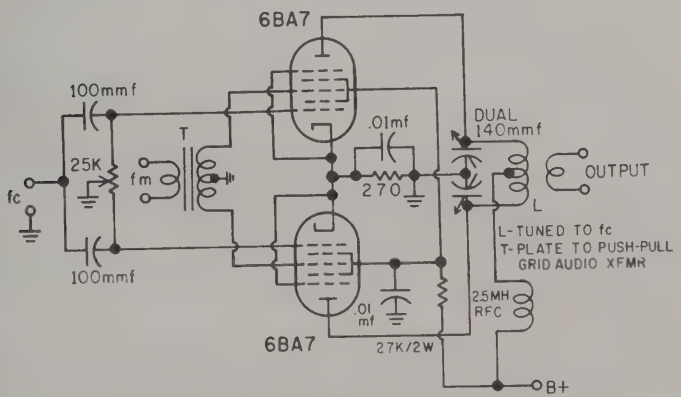


Fig. 2.5B—The pentagrid balanced modulator. This circuit is seldom used because of the additional cost.

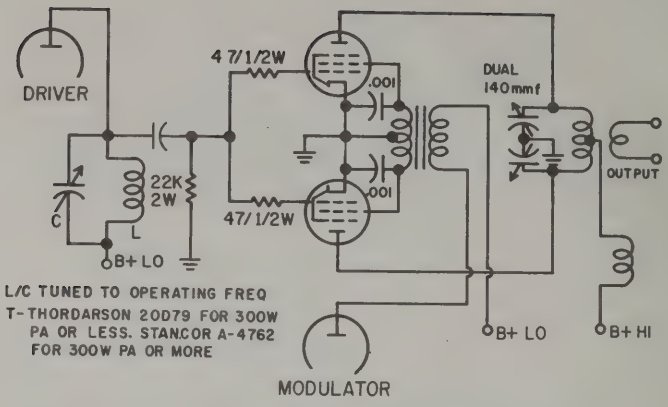


Fig. 2.5C—Common balanced modulator configuration associated with double sideband transmitters. Note that no dc is applied to the screens.

The Balanced Modulator for Double Sideband

It is possible to construct or convert a final amplifier stage to a balanced modulator. One of the advantages of DSB is the ease with which this may be accomplished. A typical double sideband power amplifier is illustrated in fig. 2.5-C. You will note that it is almost identical to the balanced modulator circuit of fig. 2.5-A, except for one very important change. The screen grids are returned to ground through the modulation transformer (T) rather than to B plus, as is the usual case. The balanced modulator action is identical to that in fig. 2.5-A, but the tubes conduct during an applied positive half cycle of audio.

Any tetrode or pentode will function in this circuit. The peak power output which can be obtained from a pair of tubes in this service may be estimated by taking the carrier output from the manufacturers specifications and multiplying by four. For example, if a pair of 6146's are to be used, we find that one tube will deliver 52 watt of carrier output in class C telephony service at 600 volts (E_p). If the plate voltage is increased to 1200 volts, it is possible to obtain 208 watts peak power output. The lack of screen voltage keeps the plate dissipation well with-in the ratings even at double plate voltage!

The "Ultimate" Balanced Modulator

Many of the balanced modulators described in this chapter have serious shortcomings that result in less than optimum performance. Often the circuit must be made rather complicated to circumvent the problem. The load presented to the signal sources by the balanced modulator varies widely. In addition, there is usually a good deal of coupling between the audio and carrier source. The use of low impedance circuitry will "tame" the load variations, but necessitates impedance matching components.

A tube designed for color television service, as a balanced *demodulator*, makes an ideal balanced *modulator* for Amateur transmitters. The 6AR8 "sheet beam" switching tube cures many of the ills associated with balanced modulators. The tube is rather unorthodox in construction and operation. Electrons leaving the cathode are formed into a "solid" sheet. Electrons pass through the first grid (pin 6) and are modulated by the carrier. A second grid serves to isolate the carrier section from

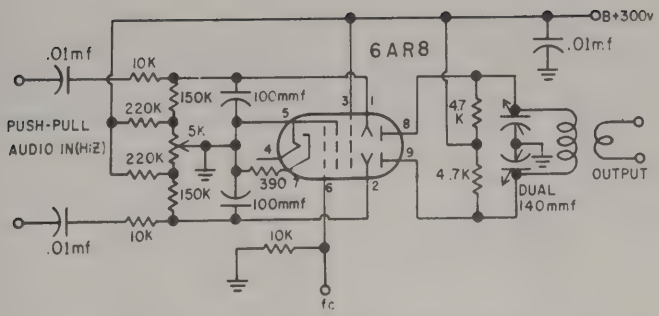


Fig. 2.5D—The "ultimate" balanced modulator employs a sheet beam switching tube. Because of its many advantages (see text), it should be popular with Amateurs and manufacturers.

the audio section. Note that because of the shortage of tube pins this grid is common connected to one side of the filament (pin 5) and this pin *must* be grounded. The sheet beam continues on thru the accelerator region (Pin 3) and is "speeded up". A high impedance source of audio is connected to the deflection electrodes and serves to switch the beam between the two plates. With no audio applied, an equal amount of r-f (modulated electron beam) will reach each plate and cancel in the push-pull tank circuit. A small amount of d.c. is applied to the electrodes and serves to "center" the beam, and balance out the carrier.

The circuit shown in Fig. 2.5-D was designed by the author and is an adaption of the circuit described by W. I. Orr (W6SAI) in the July 56 issue of CQ Magazine. The main advantage of the 6AR8, when used as a balanced modulator, is that it works well when the input and output circuits are of relatively high impedance. One-half of a 12AX7 may be employed as an audio phase inverter to supply push-pull signals to the deflection electrodes, rather than a push-pull audio transformer. Usual values of plate and cathode load resistors may be employed in the phase inverter circuit, without fear of excessive circuit loading by the balanced modulator stage. It is not necessary to use a low-mu triode (with the accompanying low values of plate and cathode load resistance) such as is necessary when driving a metallic diode modulator.

In addition, there is very little reaction between the mixing oscillator and the balanced modulator. A balance capacitor (see section 2.7) is seldom necessary. Many balanced modulators impose a non-linear load upon the mixing oscillator. If this oscillator is a variable frequency device, frequency modulation will occur during the mixing process. This action produces a peculiar low frequency "growl" on the SSB signal which is most annoying, and difficult to cure. This effect is entirely absent when the mixing oscillator is coupled to the 6AR8 modulator.

To obtain conditions of balance, the deflection electrodes are "d.c. balanced" with a potentiometer. This potentiometer is at r-f ground and may be located at some point remote from the rest of the circuitry. By adjusting this control, more than 50 db of carrier suppression may be obtained.

2.6 VARISTOR BALANCED MODULATORS

The term "varistor" may be new to some readers. To dispel any confusion that might exist, a varistor is a group of usually two or four identical rectifying units.

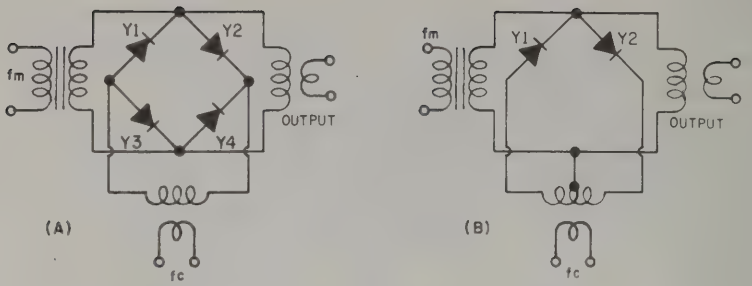


Fig. 2.6A—Basic two and four varistor shunt modulators.

The individual units may be the familiar 1N34 germanium diode, a copper oxide rectifier, a silicon diode or even a selenium rectifier. The "varistor" group may be sealed in a shield can or glass envelope to protect them from damage.

The individual rectifier units of a varistor group are usually selected so that their characteristics are as closely matched to the other rectifier units in the varistor as possible. Thus a type 1N35 varistor is nothing more than a pair of carefully selected 1N34 germanium diodes.

There are three basic types of circuits in which varistors can be used as balanced modulators. These are: shunt, series, or the ring type. In order to achieve balancing action, the varistors can be connected in various ways using either two or four diodes in the varistor unit. It is the various combinations of circuitry and varistor configurations that are of interest to us.

It is interesting to note that in the usual balanced modulator circuit the ratio of r-f voltage to audio voltage is of the order of 10/1, or 20/1. That is, the carrier voltage must be 10 to 20 times the amplitude of the modulating signal for best linearity. Here, the situation is reversed. The modulating signal is just about 20 times greater than the carrier signal. Since the two signals are applied to different electrodes, of the modulator tube, the old 10/1 or 20/1 ratio does not hold true. This also means that the r-f signal source can be operated at a lower level for increased stability.

Shunt Connected Varistor Modulators

Fig. 2.6-A shows the two possible connections using a four unit varistor and a two unit varistor. Fig. 2.6-A(a), the four unit varistor in shunt connection, has the advantage of requiring no center tapped windings for the carrier supply, f_c, for the modulation voltage, f_m, or for the output transformer. It has the disadvantage of being awkward to balance. There is no really convenient place to put a balancing resistor. The situation boils down to putting a small value of variable resistance in series with one of the diodes. Finding the right diode is a matter of cut and try.

Fig. 2-6-A(b) shows the basic circuit of the two diode shunt varistor which produces a balancing action for f_c if the voltages either side of the r-f transformer center tap are equal. This is usually not the case in the practical circuit so the schematic of Fig. 2.6-B was evolved. The balancing function is now performed by the 1000 ohm potentiometer. The need for a center tapped r-f winding is now eliminated.

There is another matter that must be considered in the shunt connected circuit in order to realize efficient operation. This is the matter of the input and output impedances. The impedance of the f_m source (the audio transformer secondary) must be low at the audio frequencies and must appear as a high impedance at the carrier frequency. The impedance of the output circuit should be a low impedance at the

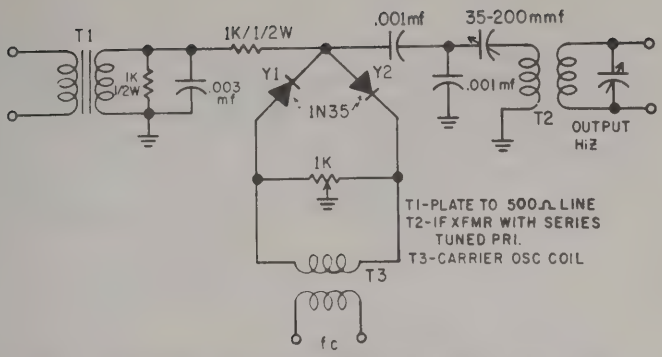


Fig. 2.6B—Practical adaptation of the circuit shown in Fig. 2.6A(a).

carrier frequency and a high impedance at the audio frequencies. The latter condition is easy to realize by the use of a series tuned circuit in the output, and resonant at the desired r-f frequency. Accomplishing the impedance condition for the audio frequency input circuit is a matter of compromise. The transformer winding has an impedance of 500 ohms at audio frequencies and the associated R/C network helps to bring about the desired conditions at the higher carrier frequency. This matter of having the correct impedances available for the input and output circuits often makes the use of the series type of modulator more desirable.

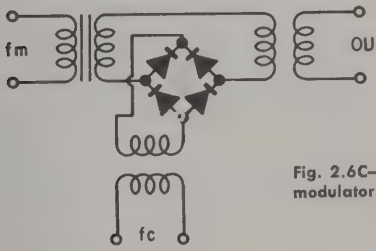


Fig. 2.6C—Basic circuit for the series balanced modulator using a four unit varistor.

The Series Type Varistor Modulator

We now come to the series connected varistor modulator covered in fig. 2.6-C. This is the four varistor circuit in which the carrier, f_c, cannot appear in the output circuit when the varistor is perfectly balanced. Application of modulating signal, f_m, disturbs the balanced condition and upper and lower sideband signals about the carrier frequency are generated. Balancing this four diode varistor series circuit is as hard to achieve as in the shunt circuit part. For this reason, the circuit of fig. 2.6-D is preferred. The potentiometer is used as a balancing control.

The input and output impedances of the modulator must now be considered. The impedance of the audio source must be high at audio frequencies and low at the r-f carrier frequency. This is accomplished by using a high impedance secondary on the audio transformer and shunting it with a capacitor which appears as a low

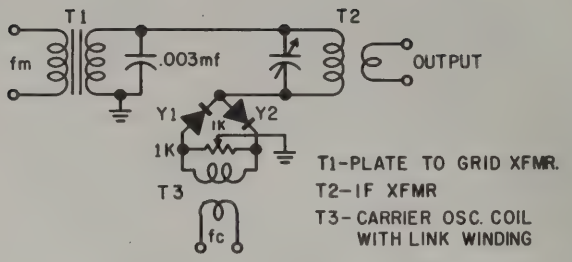


Fig. 2.6D—The series type balanced modulator employing a varistor arrangement.

reactance at the r-f carrier frequency. The reactance is negligible (with respect to the transformer secondary impedance) at the highest audio frequency used, i.e. 3000 cps. Another impedance condition that must be met is that the output impedance of the modulator must be high at the r-f output frequency and low at the audio frequencies. This condition is most easily accomplished by a parallel tuned circuit. This circuit can be just an ordinary L/C tuned circuit or it can be the input circuit of a filter whose input impedance is high in the filter passband and low outside the passband limits.

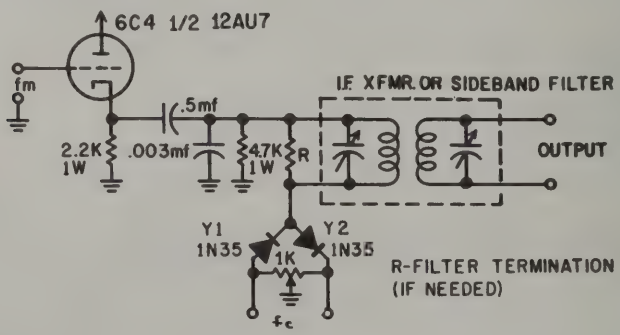


Fig. 2.6E—Modified series balanced modulator.

The circuit shown in fig. 2.6-E could be used with a Collins mechanical filter which has an input impedance of 16,000 ohms in the passband region. The input circuit is a parallel tuned circuit which will fall to a low value of impedance at audio frequencies. The cathode follower stage was substituted for the input transformer of fig. 2.6-D. The cathode follower load resistor (2.2K) is high impedance when compared to the filter impedance for the audio frequencies. The shunting capacitor (.003) is a low reactance at 455 kc when compared to the 16,000 ohms filter characteristic impedance, yet has negligible shunting effect at 3000 cps. The blocking capacitor (.5 mf.) prevents the cathode voltage from biasing the diodes and disturbing operation.

Ring Modulation

Fig. 2.6-F is the classic circuit used by the telephone companies for the land line SSB generators. This same circuit has been used by many hams using filters in the 1.0 to 50 kc. region. The input and output impedances of this circuit are low — about 600 ohms. The balancing potentiometer is placed between the two halves of the winding of the output transformer primary. The impedance of the carrier voltage feed is very low and is most easily matched by using a link winding on the carrier voltage supply.

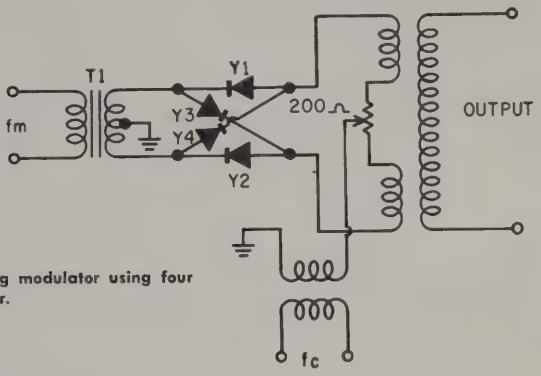


Fig. 2.6F—The telephone company ring modulator using four copper oxide rectifiers as a varistor.

The Collins Balanced Modulator

An interesting variation of this circuit is used by The Collins Radio Company in their single sideband transmitters. The circuit may be used at low frequencies ranging up to 500 kc, and is shown in fig. 2.6-G. The 250 kc carrier frequency, because of equal inputs is balanced out, or suppressed. The 100 ohm potentiometer and 5-25 mmf capacitor provides vernier adjustments for the balance circuits to take care of inherent component variations. The 1N67A diode rectifiers are made up in matched sets at the factory.

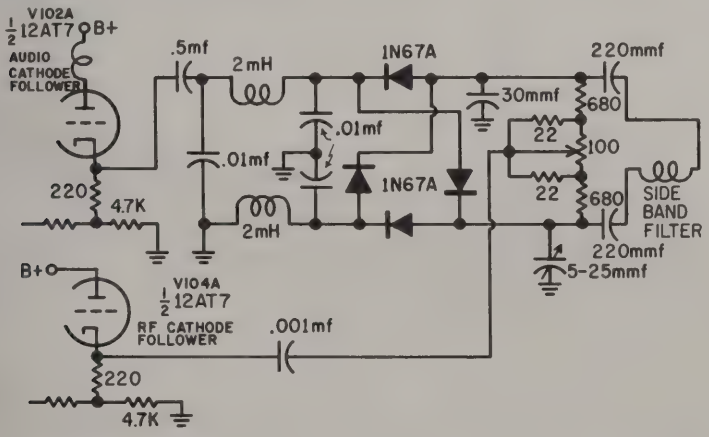


Fig. 2.6G—A practical application of the ring modulator. This circuit is used to create the sidebands and supress the carrier in the Collins KWM-1 and KWS-1.

The Modified Ring Modulator

The circuit shown in fig. 2.6-H is one form of modified ring modulator that has proven very useful in SSB exciters. This basic circuit is used in the "SSB-Jr." described by W2KUJ in *GE Ham News* (Vol. 5, No. 6) November-December 1950 issue. This circuit is also used in the current Central Electronics and Lakeshore exciters. The two signals are connected in series to the common feed point of Y1 and Y2 which is the arm of the balance potentiometer. The secondary of T1 is a low impedance winding — usually a voice coil winding — and L is a link coil wound on the carrier oscillator coil. The push-pull tank circuit must resonate at the frequency of the desired output.

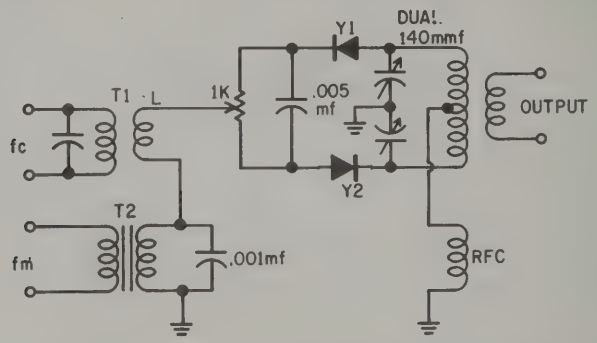


Fig. 2.6H—Commercial application of the modified ring modulator. It is used in the Central Electronics, Lakeshore and Johnson single sideband equipment

A Transistor Balanced Modulator

Two transistors can be employed in a balanced modulator configuration as shown in fig. 2.6-1. The transistors need not be matched and the types specified

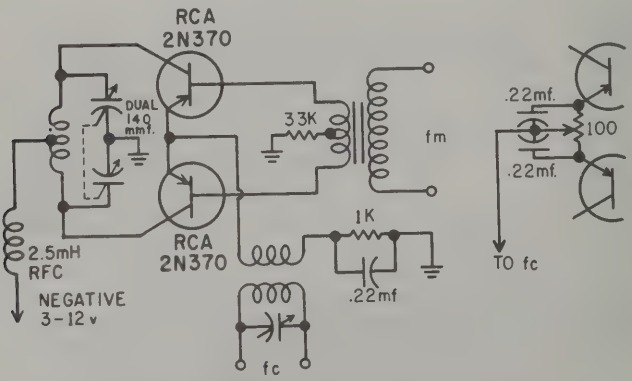


Fig. 2.6I—Transistors can also be used for balanced modulators. The optional balance circuit (shown at right) may be substituted if additional carrier balance is required.

will usually balance out quite well. Bias for the modulators is supplied by rectification of the carrier which is fed to the common emitters. The positive half cycles of rf will cause base current to flow. This current develops a voltage across the 1K emitter resistor thereby providing bias for the units. The signal f_c is supplied to both transistors in phase. Since a push-pull tank is used in the output, carrier cancellation will occur. The modulation, f_m, is connected to each base 180° out of phase. When modulation is applied one transistor will conduct more, while the other will conduct less. During the next half cycle, the condition reverses.

Because of the push-pull configuration the transistorized balanced modulator is not particularly temperature or level sensitive. With the units specified, the level of f_c should be 10- 25 milliwatts. Excessive drive will change the transistor characteristics or even cause destruction of the units. If the "last ounce" of carrier suppression is needed, a 100 ohm balance potentiometer may be inserted between the two emitters, with the carrier, f_c, connected to the arm. This will allow balance of all 2N370 types.

2.7 CAPACITIVE BALANCE

In the discussion thus far, resistive balance has been covered, but no mention has been made of correction for unequal tube and circuit stray capacities. Quite frequently no capacitive balance circuit is needed. The resistive balancing potentiometer will in most instances give quite satisfactory balancing action.

If capacitive balance is also needed, a small variable capacitor of the air padder or compression mica type should be added to either one of the two possible points in the balanced modulator circuit. Finding the correct point to connect the correcting capacitor is a matter of experimentation. This added capacitor should be connected between ground and the point in question. If the capacitive unbalance is too great for the variable range available on the correcting capacitor an additional fixed mica capacitor should be added to bring the balance point within range.

The balance capacitor should be connected between ground and one of the grids of fig. 2.1-B. A 100 mmf. compression mica capacitor is usually adequate. Either plate of fig. 2.2-A is used for balancing. The balance capacitor should be placed from one grid to ground in fig. 2.3-A. One control grid of fig. 2.5-A is used while one of the oscillator injection grids (grid #1) of fig. 2.5-B is the correct place for the balance capacitor.

The series and shunt types of varistor balanced modulators (shown in figs. 2.6-A, 2.6-B, 2.6-C and 2.6-D) are usually handled by connecting the balance capacitor to one side or the other of the carrier source. If an isolation transformer is used between the carrier oscillator and the varistor, the balance capacitor should be connected to one of the secondary windings.

The ring modulator of fig. 2.6-F is best corrected for capacitive unbalance by placing the small balancing capacitor from one of the outer ends of the output transformer primary to ground. The modified ring modulator of fig. 2.6-H is also balanced by connecting the balancing capacitor from one side of the tuned output coil to ground.

2.8 SPURIOUS SIGNAL GENERATION

Use of the balanced modulators described in this chapter has been restricted to modulating an r-f carrier frequency signal with a speech-frequency signal (0 to 3000 cps). Practically any of the circuits described can be used to convert a single sideband signal of one frequency to another frequency where it can be used. For

example, if a filter type exciter is used and the sideband is generated at 455 kc, we must convert this into the 3.9 mc band so that we may make use of it. The first thought that might come to mind would be to use an ordinary pentagrid mixer tube and not use a balanced modulator at all. This is often the simplest solution. There are occasionally instances where spurious signals are generated in the mixing process. The presence of these signals will often make the more complex balanced modulator circuits attractive.

The spurious signals are formed in the same manner as intermodulation distortion products in a linear amplifier. The output tuned circuit for a frequency converter will not generally be tuned to either of the two input signal frequencies f_1 or f_2 but will be tuned to either $f_1 + f_2$ or $f_1 - f_2$ so that a new frequency f_3 is formed.

The spurious signals we are concerned with will be those that fall near f_3 in the frequency spectrum. The proximity of danger will be dependent upon the selectivity of the tuned circuits that follow. The actual generation takes place because of the mixing action that will occur between harmonics of f_1 and the harmonics of f_2. For example, the 4th harmonic of f_1 might mix with the 7th harmonic of f_2 to produce a "difference" mixture that falls near f_3 in frequency. Even if the wave shapes of both f_1 and f_2 signals are free of harmonic content, the mixing process generates harmonic frequency signals that will cause trouble.

The ordinary variety of mixer stage, where both f_1 and f_2 signals are fed into the tube in a "single-ended" arrangement and the output circuit is also single, is probably the worst offender. This circuit will form both the odd and even harmonics of both input signals, f_1 and f_2. The picture may appear darker than really need be, for the situation can be helped in the single-ended mixer circuit by keeping a high ratio between the amplitude of the heterodyning oscillator signal and that of the SSB signal to be converted. The desirable ratio might be as high as 100 to 1.

If, however, trouble is encountered, one of the various forms of balanced modulators should be tried. Before starting, it should be determined by some scratch pad calculations, just what harmonics of the f_1 and f_2 signals are causing trouble. If it is an even harmonic of f_1 mixing with an odd harmonic of f_2 the problem is to eliminate either one of the offending harmonics being generated. It is not necessary to eliminate both. If one is eliminated the spurious mixture product will be eliminated.

The best "all-around" solution would be to use the push-pull type of balanced modulator shown in fig. 2.1-B. The parallel feed of the heterodyning oscillator signal, f_{osc}, and the push-pull connection of the output circuit will tend to cancel all harmonics of the f_1 signal. This is fortunate if the related circuits to be used, permit the configuration shown. If it is not possible to use such an arrangement, an intelligent choice must be made among the other types of circuits.

Should you plan to design and construct a sideband exciter, it is recommended that you try to obtain a copy of "The Mix-Selector Chart" that was published in Vol. 11, No. 6 (November-December, 1956) issue of *GE Ham News*. The chart and accompanying explanation shows graphically, the frequencies at which beat notes or harmonics can cause trouble.

2.9 CARRIER GENERATORS

Generation of the carrier is an "orphan" subject that might well go in any of a number of other chapters. However, it is most closely associated with the balanced modulator. A carrier must be created initially so that it may perform its "duty" of creating sidebands. Once this end has been accomplished, the carrier has outlived its usefulness and it is suppressed in the balanced modulator. Actually, as you know, both these functions are accomplished simultaneously in the balanced modulator stage.

Stated simply, the carrier generator is nothing more than a radio frequency oscillator. The oscillator may be either crystal controlled or variable in frequency. In addition, the carrier may be multiplied to the ultimately used frequency so long as this process occurs before the balanced modulator stage that mixes the carrier and audio. Once it "leaves" this stage it *must* be heterodyned to the operating frequency.

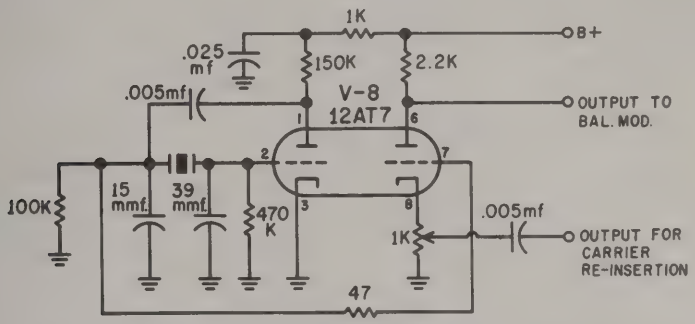


Fig. 2.9A—Simplified circuit of the Eldico SSB-100F carrier generator. This circuit works extremely well with FT-241 surplus crystals.

It is the application, then, that brings about the term carrier generator. As an example, many Amateur single sideband transmitters (filter type) generate the carrier at a low frequency where filters are not quite so tricky to work with. The Eldico SSB-100F, for instance, generates a carrier on 413 kc. The circuit is a Pierce, modified for proper impedance matching of the crystal, and is shown in fig. 2.9-A. The oscillator stage is untuned and the output is coupled to a so-called "paraphase" amplifier. The 413 kc energy is available at either the plate or the cathode. The plate circuit signal is used to drive the balanced modulator. A potentiometer in the cathode circuit can be used to select the desired amount of carrier should AM operation be desired. This "reinserted carrier" is applied to the stage following the sideband filter.

This circuit will also work very well with the war surplus FT-241 crystals although it may be necessary to add capacity on each side of the crystal to insure stable oscillation. The input capacity of the oscillator triode will have a pronounced effect on the frequency of oscillation. A small trimmer capacitor substituted for the 39 mmf. capacitor may be used to "rubber" the crystal frequency.

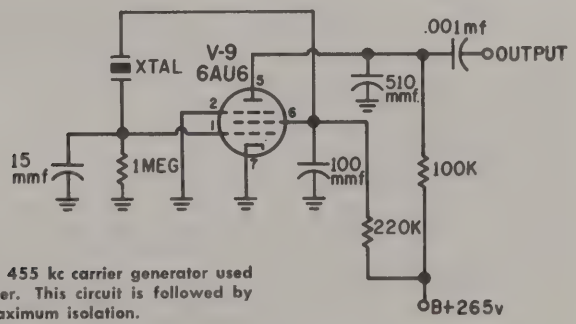


Fig. 2.9B—The electron coupled 455 kc carrier generator used in the Collins KWM-1 transceiver. This circuit is followed by a buffer/amplifier to provide maximum isolation.

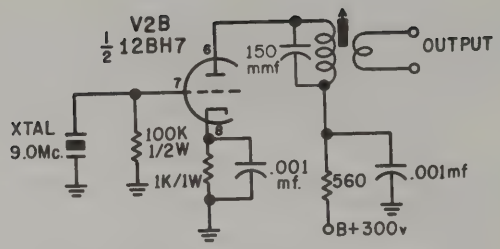


Fig. 2.9C—Diagram of the 9 mc carrier generator used in the Central Electronics Multiphase exciters.

Another circuit that is suitable for use with low frequency war surplus crystals is shown in fig. 2.9-B. This circuit is used in the Collins KWM-1 for carrier generation on 455 kc, and incidentally for the 100 kc crystal calibrator. This is an electron coupled type of oscillator, with feedback between the screen and control grids. This portion of the tube functions as a triode and the screen acts like a "plate with holes in it". A small amount of the signal varying the electron stream is "picked off" by the screen grid and fed to the control grid via the crystal. The electron stream also passes through the grid to the plate and output circuit. A rather large capacitor across the output circuit tends to minimize frequency shift due to load variations. Although there is no electrical connection between the oscillator and load, this circuit is usually followed by a pentode buffer or a triode cathode follower. As stated earlier, the balanced modulator usually represents a "wild" load and exhibits wide impedance swings. The pentode serves to isolate these variations while the cathode follower minimizes the effect by providing a low source impedance.

The frequency of oscillation may be varied over a narrow range by inserting a *small* inductance in series with the crystal or by substituting a 8-50 mmfd. trimmer in place of the 15 mmf fixed capacitor. Departing from the circuit and values shown will affect the stability adversely, however.

The generation of the carrier need not be accomplished at a frequency lower than the ultimate operating frequency, either. The Central Electronics "Multiphase" series of exciters use a 9 mc carrier generator. After the carrier is applied to the balanced modulator, and the sideband created, the energy is heterodyned to the ultimate operating frequency. The circuit for this oscillator is shown in fig. 2.9-C. Although the frequency of oscillation is not critical, the plate inductance will shift the frequency of oscillation slightly. A complete schematic diagram of the Central Electronics 10A exciter and the theory of operation is presented in the chapter devoted to phasing exciters.

It is also possible to generate the carrier and sidebands on the operating frequency. This system is used in the sideband adapters of Barker and Williamson (Model 51-SB) and Heath Company (Model SB-10). An adapter differs from an exciter in that the carrier is generated outside the device (usually in companion AM transmitter). With either of these adapters, the carrier is removed from the power amplifier input and routed to the accessory unit. The adapter "converts" the carrier into a sideband signal and returns it to the power amplifier input. Use of these adapters makes it quite simple to modify an AM transmitter to do "double duty". Circuitry and data on this type of adapter wil also be found in the phasing method chapter.

REFERENCES

Berry, "The Series Balanced Modulator," QST, September, 1952, p. 46.
Wright, "Carrier Generators for SSB Reception," QST, December, 1952, p. 35.
Orr, "An Unusual SSB Modulator," CQ, July, 1956, p. 24.
Goodman, "SSB Balanced Modulators," QST, February, 1957, p. 42.

3

The Filter System

BY THIS time you should be aware that there are two principal systems for generating a single sideband, suppressed carrier signal. The phasing system combines audio and r-f energy, of the proper polarities, to cancel one sideband, while at the same time reinforcing the other. This chapter deals with the creation of the sideband by filtering out the unwanted sideband.

Most Amateurs have their own opinion as to which system is preferred. It is the authors' intention to "steer clear" of the melee by advocating a "no sideband, suppressed carrier system". Actually, each system has its own merits as indicated by the nearly equal balance of commercial equipment. The Central Electronics "Multiphase" line, Lakeshore, Johnson, Barker & Williamson, and Heath units are of the phasing type. The Hallicrafters, Collins, Cosmophone, Eldico, and Technical Material Corporation devices are of the filter type. Either system is capable of producing a fine signal with proper construction and alignment techniques. Conversely, sloppy workmanship will yield a sloppy signal with either system.

3.1 FILTER METHODS AND DESIGNS

As we saw in Chapter I, an AM signal is composed of a carrier, which has no bandwidth, and two sidebands. The sidebands are like the carrier, in that they contain r-f energy, but are distributed on either side of the carrier in the spectrum. The sideband components corresponding to the lower audio modulating frequencies are close to the carrier, and the higher audio frequencies are proportionately farther away from the carrier. The two sidebands are identical and contain the same intelligence. The filter type of exciter must separate these two sidebands by passing one and attenuating the other. The matter of carrier suppression is secondary and can be accomplished with ease in the balanced modulator.

It is not feasible, from a practical standpoint, to accomplish the filtering process at the final operating frequency for a number of reasons. First, finding circuit components to separate the sidebands, whether coils, capacitors, or crystals, would be an almost impossible job. It would require almost an infinite number of components! If it was desired to change frequency a new filter would be required, or all the circuit components would have to be re-tuned for a new set of conditions. There is, however, an easier way to eliminate the unwanted sideband. The single sideband signal can be generated at some lower fixed frequency and then heterodyned up to the desired frequency with one or more mixers. The selection of mixer circuit will be dictated by the possibility of harmonics or beats that might cause spurious signals to be radiated. The mixing procedure solves both the preceding problems in that it makes the filter requirements simpler. A variable frequency oscillator can be used as a heterodyning oscillator to place the single sideband signal anywhere in a particular ham band.

The Nature of Filters

The operation of a filter and the terms associated with this device might present some confusion. By definition, a filter is a device that presents a high impedance to unwanted frequencies, but offers little resistance or attenuation to the desired frequencies. The amount of attenuation it presents to the desired frequencies is termed *insertion loss*. The frequencies passed by the filter are logically called the filter *pass band*. At some frequency the attenuation of the filter will start to increase rapidly and the plot of this frequency versus attenuation is called the *skirt selectivity*. A plot of the entire filter response would be referred to as the *characteristic curve*. The *flat-top* is that portion of the curve where minimum attenuation occurs. The *bandwidth* of a filter can be a rather indefinite thing. Some manufacturers specify the width of the flat-top while others specify the point on the skirt at which the attenuation causes the applied voltage to drop to half its normal value (.707). Still other manufacturers may specify the frequency at which an attenuation of 20 db occurs. The merit of a sideband filter is called the *shape factor*. The figure is the ratio of bandwidth at 6 db and 60 db points. A typical filter might be 3 kc wide at 6 db and 6 kc wide at 60 db. This is a shape factor of 2.0 and is considered extremely good.

Filters can be divided into four distinct classes, i.e. low pass, high pass, band pass and band elimination. A low pass filter will transmit frequencies up to some point and then attenuate any frequency above this point. The frequency at which the selectivity skirt drops to 6 db is referred to as the *cut-off frequency*. An example of a low pass filter is that wonderful device that will be found on most amateur transmitters to minimize television interference. The band pass might extend to 35 mc, at which point the attenuation would increase to a very high value. A companion filter for this device will be found on many television receivers to reject frequencies below the lowest television channel (54 mc). This device is termed a highpass filter. We are most concerned with the band pass type of filter. The SSB form is character-

ized by a steep skirt, 2.5 to 3.5 kc bandwidth at the 6 db point, and a high degree of attenuation on either side of the design center. The band elimination filter can be found in many ham shacks, but by a different name. The Q Multiplier and T Notch are actually band elimination filters, but in this case the "band" is a single unwanted frequency; namely the heterodyne. These four types of filters are shown in fig. 3.1-A, along with the associated characteristic curves. None of these curves are very well suited for eliminating the unwanted sideband, although the band pass filter curve is the closest. It is necessary to use more complicated forms than the coil and capacitor. The band pass filters currently being used are the inductive filter, the crystal filter and the mechanical filter. Since all of these types are suitable for eliminating the unwanted sideband, they will be examined in detail.

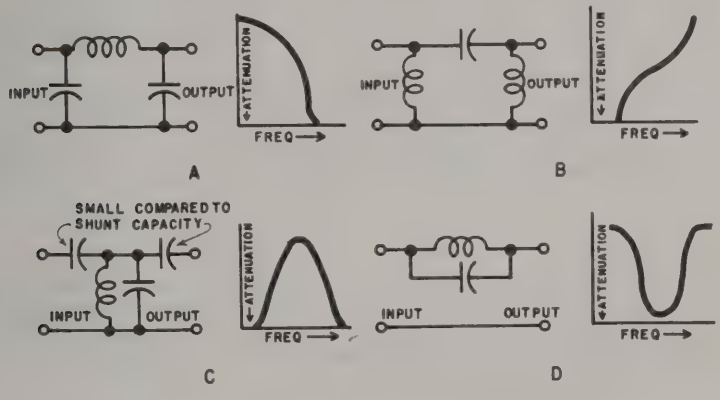


Fig. 3.1A—Four basic forms of filters referred to in the text. The configuration produced by the bandpass filter is most suitable for eliminating the unwanted sideband.

Filter Exciter Designs

The sideband filter must have certain characteristics to be acceptable for use. The filter, whether band pass, high pass or low pass in nature, must be capable of separating the sideband components close to the carrier on the high frequency side from those close to the carrier on the low frequency side. This dictates a sharp drop off in the filter response characteristic, or a good skirt response. We cannot obtain practical filters with a vertical drop off characteristics, so we must be content with an attenuation of from 30 to 60 db in about 600 cps. The situation can be improved by moving the carrier frequency part way down the slope of the filter. Two things are accomplished by doing this: The useless, power wasting low speech frequencies below 300 cycles are attenuated. The filter becomes more effective on the portion of the undesired sideband near the carrier. A good rule of thumb to follow in placing the carrier on the slope of the filter is to put it about 20 db down on the skirt. This should result in the transmission of few frequencies below 300 cps. If the carrier is so placed and the speech response is not up to maximum until you reach a frequency of 700 or 800 cps, the filter skirt is not steep enough and measures must be taken to correct this. The commercial standard for a good sideband filter is 80 db drop in the filter skirt in 1 kc! This is a bit more than required for amateur use, but it will give you something to shoot for!

The high frequency response of the filter characteristic and speech amplifier need only be enough for good speech intelligibility. This means that frequencies above 3000 cps should be attenuated, first of all, to conserve spectrum space, and to result in a pleasant sounding signal. Since we have eliminated the low frequency speech components, cropping the high frequencies will produce a balanced sounding signal that will be more intelligible.

Heterodyning

Since filtering of the unwanted sideband cannot be accomplished on the operating frequency in a practical manner, it must be done on a lower frequency and heterodyned up to the desired band. This concept is basic to all filter type exciters. A simple filter exciter for 75 meters is shown in block diagram form in fig. 3.1-B. You will note that an upper sideband filter (passes upper sideband) is used. When this 450 kc upper sideband signal has been heterodyned to the 4.0 mc band it becomes a lower sideband and signal. How come? Using a hardwood calculator (pencil), figure the simple mathematics involved. Assume 450 kc as a suppressed carrier frequency and 452 kc as the frequency of an upper sideband resulting from a 2000 cycle input tone. It is the difference mixture of this frequency and the vfo signal that produces the operating frequency. Subtracting 450 kc from the 4350 vfo frequency results in an operating frequency of 3900 kc. However, subtracting the 452 kc sideband from the vfo frequency produces a sideband frequency of 3898.

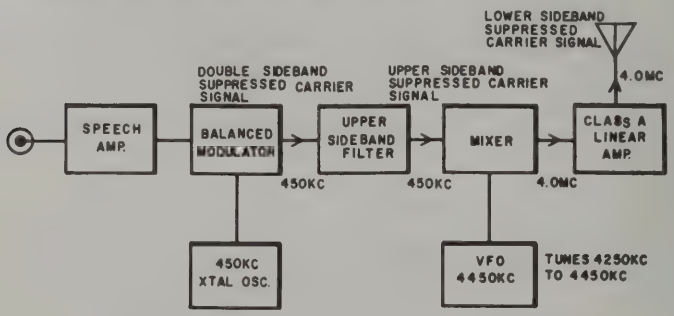


Fig. 3.1B—Block diagram of a simple filter type SSB exciter.

Since this is on the lower side of the suppressed carrier frequency, it is called a lower sideband. This effect is called *sideband inversion*. It will always result in a mixer stage when the *difference mixture* is utilized. If it was desired to transmit an upper sideband signal on 75 meters, it could be accomplished in one of two ways. A lower sideband signal could be generated in the low frequency stage (455 kc) or the heterodyning oscillator could be moved to *below* the operating frequency so that a sum mixture would result.

A very good example of the various heterodyning processes is shown in fig. 3.1-C. This is the block diagram for the Hallicrafters HT-32 exciter. Elimination of the unwanted sideband occurs at 5 mc, using a high frequency crystal filter. The sideband is then heterodyned to 9.0 mc with either a sum or difference oscillator (for sideband selection). For all bands except 75 and 20 meters, the signal is again heterodyned to 5 mc away from the operating frequency. A variable frequency oscillator operating between 5.0 and 5.5 mc finally heterodynes the sideband to the desired frequency. Although this exciter employs several conversion oscillators, spurious signals are non-existent.

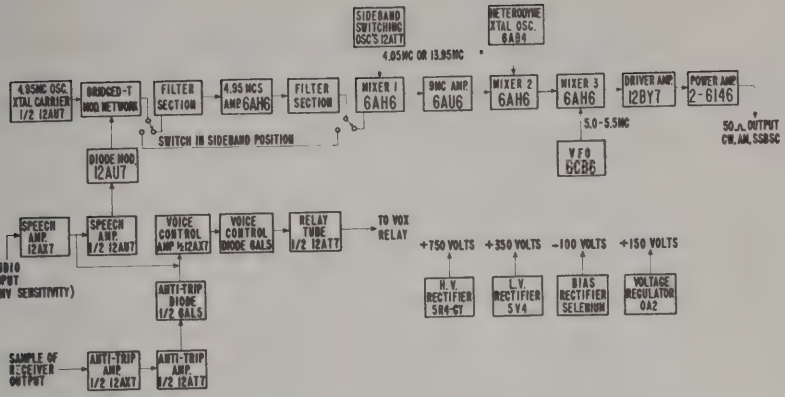
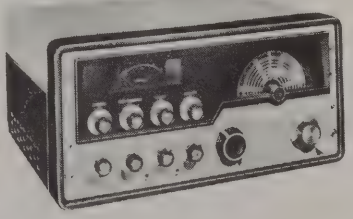


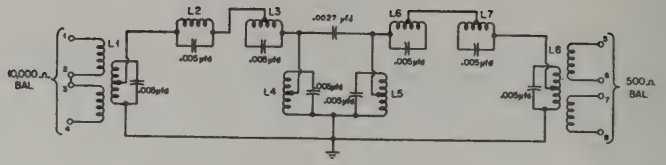
Fig. 3.1C—Block diagram of the Hallicrafter HT-32 exciter. The sideband is generated on 5 mc and the first mixing places it on 9 mc. It is then heterodyned to plus or minus 5 mc from the operating frequency, except on 75 and 20 meters. The difference between the second conversion and the vfo determines the operating frequency on 40, 15, 11 and 10 meters.

3.2 THE INDUCTIVE FILTER

The simple LC tank circuit does not have the necessary characteristics for a good sideband filter. It does not have steep sides, a flat top or a narrow bandwidth. Hence, it is necessary to use a complex arrangement of LC circuits. The resonant circuits are usually designed for a very low radio frequency (10- 50 kc) since the tuned circuit bandwidth becomes smaller as frequency is decreased. An example of such a filter is shown in fig. 3.2-A and is designed for operation between 17 and 20 kc. Early types of filters were very similar to this configuration. Direct heterodyning into Amateur bands from this low frequency is not practical for when two signals are mixed, the output products are the original signals and the sum and difference of these original signals. For example, if a 25 kc signal were to be heterodyned directly to 4.0 mc the tuned circuits would not allow the original 25 kc signal to appear in the output. But the mixing oscillator near 4.0 mc would be present and both the sum and difference of the two signals would appear in the output. Of course, if a balanced mixer circuit were used, the original mixing oscillator could be balanced out, but two sideband, 25 kc above and below this frequency would still be present. Transmitting two sidebands (50 kc apart) is undesireable to say the least! The single tuned circuit is just not sharp enough to eliminate one sideband without eliminating the other. Thus, it is necessary to first heterodyne the 25 kc signal up to an intermediate frequency such as 455 kc. This will allow



The Hallicrafters HT-32 is the first commercial transmitter to employ a high frequency crystal filter. The unwanted sideband is filtered out at 5 mc and the wanted sideband is heterodyned to the operating frequency.



Winding	Approx. Total Ind. mhy.	Ind. to Tap mhy.	Test Capacity ufd.	Test Freq. - Kc.
L_1 Primary 10,000	16.5 (total)	(Note: - This winding is not critical)		
L_1 Secondary	14.9	4.98	.01937 for tap	18.44
L_2	10.96	no tap	—	21.50
L_3	20.25	14.90	.006798 for tap	15.814
L_4	14.4	2.408	.029883 for tap	18.80
L_5	13.9	2.65	.026235 for tap	19.20
L_6	12.06	4.13	.014524 for tap	20.50
L_7	18.4	5.13	.018366 for tap	16.585
L_8 Primary	14.9	6.6	.011289 for tap	18.44
L_8 Secondary 500	1.09 (total)	c.t.	.068120 for total winding	18.44

Fig. 3.2A—An inductance filter designed by F. M. Berry, WØMNN. The test capacity specified is used only to adjust the coil inductance to the tap. The total winding must be tuned to the test frequency by its own .005 mf mica capacitor, and then wired into the circuit. Molybdenum Permalloy toroid forms are used. The L1 and L8 forms are Arnold type P-476930 and L2 through L7 use P-284395 forms. The forms may be purchased from the Arnold Engineering Co., Marengo, Ill.

us to use normal i-f transformers to discriminate between the various products of mixing and select the desired one. Then it is necessary to heterodyne once more into the desired Amateur band making this mixing oscillator variable for "moving around" inside the band. It should be noted that even 455 kc is not too suitable for frequencies above the 40 meter band. Just as it is possible to have images on a 10 meter receiver (using a 455 kc i-f), it is possible to *transmit* images. In other words, it is not possible to heterodyne from 455 kc directly up to 10 meters unless many tuned circuits are used to separate the image from the desired frequency. The first transmitters described for Amateur use utilized low frequency tuned circuits. They were a little complex in adjustment, number of tubes and tuned circuits.

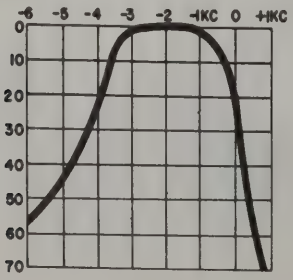


Fig. 3.2B — Frequency response characteristics of the Burnell and Co. Model S-15000 toroid sideband filter. Note that the skirt response falls off very rapidly, more than 50 db in 1 kc.

Fig. 3.2C—Block diagram of the Technical Material Corporation Model SBE-1 (Jan * AN/URA-23) transmitter. Note that it has provisions for dual channel input so that independent information may be applied to each sideband.

However, they produced a very high degree of sideband attenuation. Burnell & Company market a 50 kc toroid sideband filter (Model S-1500) that has a skirt selectivity of 60 db in 1 kc and is very popular with the purists. A characteristic curve for this filter is illustrated in fig. 3.2-B.

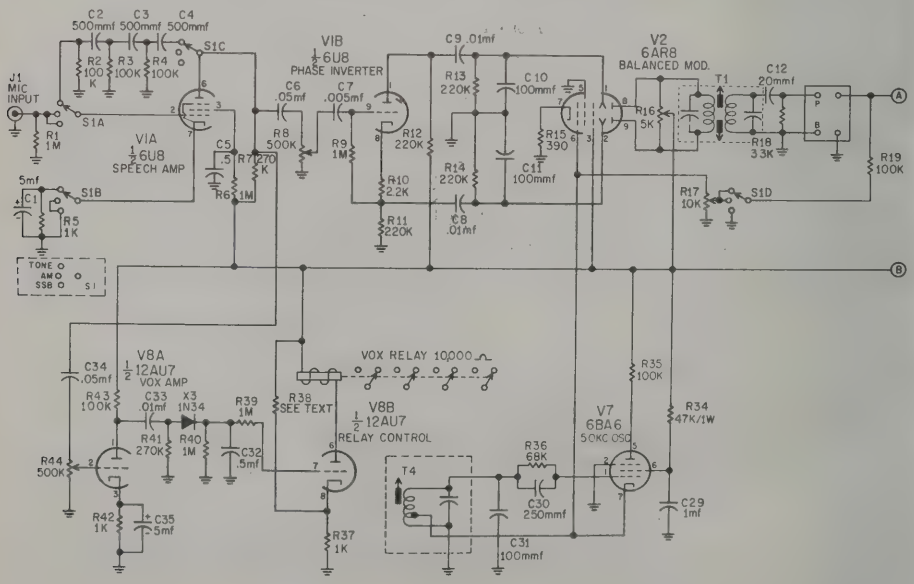
An elaborate filter transmitter for commercial, Amateur and Military use is illustrated in fig. 3.2-C. The transmitter, shown in block diagram form is the Technical Material Corporation Model SBE-1 (Military series AN/URA-23). The sideband suppression occurs on 17 kc using inductive filters. Both the 17 kc carrier generator crystal and the 287 kc crystal for the first conversion oscillator are contained in a common oven. Two additional oscillator and conversion stages are employed to cover the frequency range of 2-30 mc. The 2nd and 3rd conversion oscillator crystals are also contained in ovens. Although this transmitter uses a 6146 power amplifier, it carries an ultra conservative rating of 3 watts P.E.P.

3.3 BUILDING AN INDUCTANCE FILTER EXCITER

Low frequency filters have long been recognized as the ideal way to suppress the unwanted sideband. As the audio frequencies are approached, tuned circuit bandwidth becomes more narrow, frequency wise. Generally speaking, the filter characteristics can be more closely controlled at low radio frequencies.

Low frequency filters are seldom used in commercial equipment (for ham use) for economic reasons. The lower the filter frequency, the more conversion stages that are required to reach the desired operating frequency. Thus, higher production costs cannot compete with high frequency filter exciters. Amateurs tend to shy away from constructing such exciters because of the additional complication.

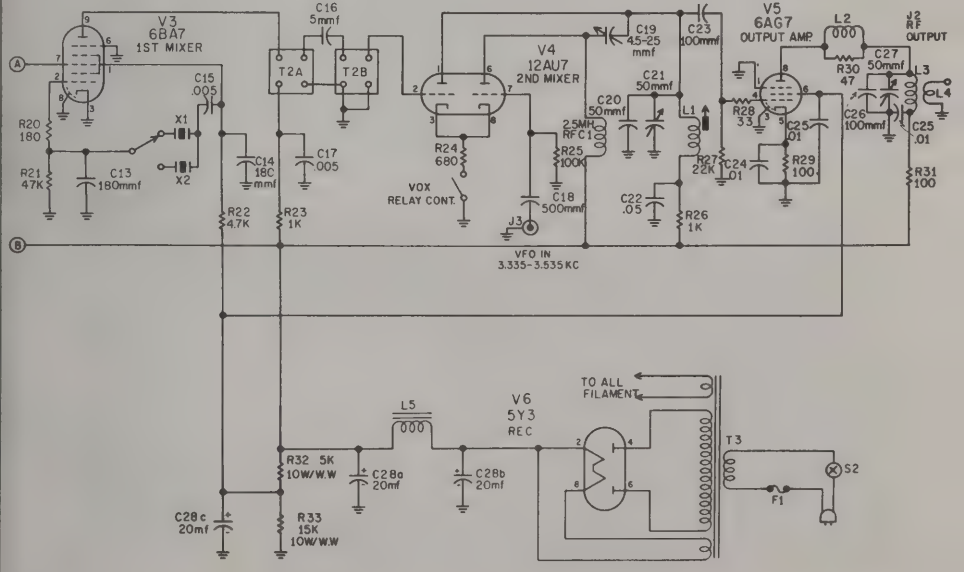
In addition, each conversion stage will be accompanied by one or more spurious signals. Unless balanced modulators are used in conjunction with selective tuned circuits, the result of these beats may reach the output.



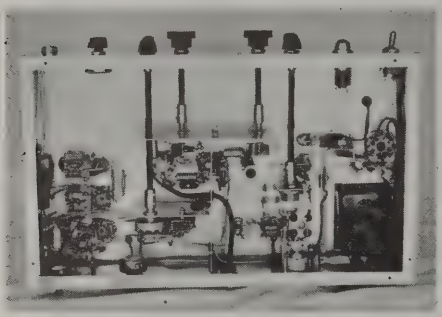
Parts List for the toroid filter exciter.

C1, C35—5 mfd., 25 volt electrolytic
C2, C3, C4—500 mmfd. mica
C5, C32—.5 mfd., 400 volt paper
C6, C22, C34—.05 mfd., 400 volt paper
C7, C15, C17—.005 mfd., 600 volt disc ceramic
C8, C9, C33—.01 mfd., paper
C10, C11, C23, C26, C31—100 mmfd. mica
C12—20 mmfd., mica
C13, C14—180 mmfd., mica
C16—5 mmfd. silver mica
C18—500 mmfd. disc ceramic
C19—4.5- 25 mmfd. rotary ceramic trimmer (Centralab 822-AZ)
C20—50 mmfd., mica
C21, C27—50 mmfd. variable capacitor
C24, C25—.01 mfd., mica
C28a, b, c—3X20 mfd., 450 volt electrolytic can
C29—.1 mfd., paper
C30—250 mmfd., mica
F1—3 ampere slo-blo fuse
FL1—50 kc. SSB filter (Burnell S-15000)
J1—Microphone connector
J2, J3—Coaxial connector (Amphenol)
L1—40 turns, #26 close-wound on ½" form (National XR-50)
L2—4 turns, #26 space-wound on R30, 47 ohm resistor.
L3—22 turns, #18 close-wound on 1¼" polystyrene coil form
L4—4 turns, #20 plastic covered wire, wound at cold end of L3
L5—filter choke, 7 henries, 140 ma. (Stancor C-1471)
R1, R6, R9, R39, R40—1 meg., ½ watt
R2, R3, R4, R19, R25, R35, R43—100K, ½ watt
R5, R23, R26, R42—1K, ½ watt
R7, R41—270K, ½ watt
R8, R44—500K potentiometer, audio taper (Centralab Radiohm B-60)
R10—2.2K, ½ watt
R11, R12, R13, R14—220K, ½ watt
R15—390 ohms, ½ watt
R16—5K potentiometer, linear taper (Allen Bradley "J")
R17—10K potentiometer, audio taper (Allen Bradley "J")
R18—33K, ½ watt
R20—180 ohms, ½ watt
R21, R34—47K, ½ watt
R22—4.7K, ½ watt
R24—680 ohms, 1 watt
R27—22K, ½ watt
R28—33 ohms, ½ watt
R29, R31—100 ohms, 1 watt
R30—47 ohms, ½ watt
R32—5K, 10 watt wire wound
R33—15K, 10 watt wire wound
R36—68K, ½ watt
R37—1K, 1 watt
R38— See text
RFC1—25 mh. r.f. choke (Miller #6302)
S1—4 pole, 3 position phenolic wafer switch (Centralab PA-1013)
S2—SPST toggle switch
T1—50 kc. i-f transformer (Miller 1898-A)
T2A, T2B—455 kc. i-f transformer (Miller 913-C1)
T3—Power transformer 360 v, 120 ma.-6.3 v. @ 3.5 a-5 v. @ 3.0 a (Stancor PM-8410)
T4—50 kc. beat frequency oscillator transformer (Miller 1898-BFO)
X1—415 kc. crystal (surplus Channel 24)
X2—515 kc. crystal (surplus Channel 78)
X3—1N34 germanium diode
Miscellaneous componets: Chassis 9" x 3" x 15" aluminum, 6U8 tube, 6AR8 tube, 2- 12AU7 tube, 6BA6 tube, 6BA7 tube, 6AG7 tube, 5Y3 tube, 2- 8 pin octal sockets, 5- 9 pin sockets, 1- 7 pin socket, 10,000 coil, four contact sensitive relay (VOX), line cord, fuse holder, and 2 xtal sockets

Fig. 3.3A—Schematic for the toroid filter exciter.



A simple inductance filter exciter is shown in fig. 3.3-A which incorporates many worthwhile features. The complete exciter uses only eight tubes and these include the VOX circuit and power supply rectifier. The heart of the exciter is the Burnell Model S-15000 toroid 50 kc sideband filter. This is an extremely popular filter for it is priced well below equivalent crystal and mechanical filters.



Liberal use of shields prevent unwanted signal pickup. One section of the shield runs directly through the center of the 6AG7 socket, shielding the input from the output. This stage refuses to oscillate even with the plate unloaded!

Circuit Details

A type 6U8 tube is used as a combination speech amplifier and audio oscillator. Function switch, S1, is shown in the TONE position. In this position, a three section phase shift network is connected between plate and control grid of VIa. Thus, signals appearing at the plate "pass through" C2, C3, and C4 and are shifted 60° in each section. The grid signal will be 180° out-of-phase with the plate signal, and since the tube produces 180° of phase shift, the stage continues to oscillate. Output from this stage (either tone or audio) drives a push-pull phase inverter.

The balanced modulator consists of a 6AR8 sheet-beam tube that was described earlier. Push-pull audio from the phase inverter is resistance coupled to the deflection electrodes, which are by-passed for r-f. Carrier energy for the balanced modulator is connected to the control grid of the 6AR8. Since the Burnell filter has an unbalanced input, it is necessary to use a balanced-to-unbalance transformer. A Miller 50 kc i-f transformer, with the primary connected in a push-pull configuration, does the job nicely. It may be necessary to reduce the size of the capacitor used to resonate the primary of this transformer, since it was noted that the slug was almost fully retracted from the coil. Normal production tolerances might not allow the coil to resonate. If this coil is tuned too far from 50 kc. it may not be possible to balance the modulator.

When the deflection electrode and the output circuit are balanced, the carrier output from this stage will be zero. When modulation is applied, however, the electrodes will have opposite polarities at any given instant. Thus, one plate will receive more energy, while the other receives less. The output signal, then, consists of a pair of sidebands with the carrier suppressed. This 50 kc DSB signal is applied to the filter for removal of the unwanted sideband. The carrier balance control, R16, will compensate for slight circuit unbalance.

The carrier oscillator consists of a type 6BA6 connected in an electron coupled circuit, adjusted to oscillate at 50 kc. The e.c.o. feature is not used, however, and the r-f energy is taken from the low impedance cathode circuit to prevent loading the oscillator. Coil T4 is a standard 50 kc bfo. unit. The grid load resistor for the balanced modulator (R17) also serves as a variable injection control for carrier insertion. In the AM or TONE position of the function switch (S1), the mixer grid load is connected to the arm of the potentiometer.

The 6BA7 serves as an oscillator and mixer. Either of two crystals can be switched into a modified Pierce circuit. The choice of crystals (Channel 24- 415 kc and Channel 78- 515 kc) will determine which sideband is transmitted. As an example, the filter passes the lower sideband, with the suppressed carrier frequency at 50 kc and the voice extends *down* to 47 kc. If this is mixed with the 415 kc oscillator signal, the heterodyne process will create a suppressed carrier frequency of 465 kc with the voice energy extending *down* to 462 kc. If the 515 kc crystal is switched into the circuit, the suppressed carrier frequency will still be 465 kc but the voice sideband will extend *up* to 468 kc.

The first mixer is followed by a 455 kc i-f filter composed of two transformers in series. This additional selectivity is necessary to reject the crystal oscillator signal which would otherwise reach the second mixer and create a spurious cw signal. It should be mentioned that the crystal is connected between the screen grid and the oscillator grid in V3. If the reader assumes that the screen grid is "a plate with holes in it" then the configuration is analogous to a Pierce circuit. An equal value capacitor, from each side of the crystal to ground, produces an "artificial centertap" inside the crystal. Thus, with 180° of phase inversion across the crystal and the same in the tube, the stage will oscillate vigorously even with sluggish crystals.

The second mixer (V4) is a 12AU7 connected in the cathode coupled mixer circuit described in section 2.4-B. It is a form of "unbalanced-balanced modulator" that requires no push-pull circuits. The vfo energy follows two paths: it appears in phase across the 680 ohm cathode resistor (R24) and 180° out of phase in the plate circuit. Mixing with the 465 kc energy occurs in the common cathode and the heterodyne *sum* appears in the tank circuit composed of L1 and C20-21. Ordinarily, the vfo would also show up in the output because of the poor selectivity of L1. However, the trimmer capacitor, C19, feeds an out-of-phase signal to the tank to bring about cancellation of the vfo energy.

The output circuit consists of a 6AG7 operating in class A, with cathode bias. With the exception of the bias system, this stage is a "carbon copy" of the Central Electronics 10A circuit that is described in section 4.8.

The power supply needs no explanation other than to say that it delivers approximately 300 volts, and 150 volts at the junction of R32 and R33.

The VOX circuit consists of a 12AU7 used as an audio amplifier and relay control tube. Audio from the plate circuit of V1a is connected to a VOX GAIN control through C34. A portion of this signal is amplified in the V8a section and applied to a diode rectifier, connected to produce a positive voltage. Hold-in time is controlled by the time constant of R40 and C32. The positive voltage is applied to the control grid of V8b, which is normally cut-off by the bias supplied from R37 and R38. The value of R38 controls the amount of cut-off current that V8b draws and its value will be determined by the VOX relay that is used. For a 10,000 relay a value of 68K, 2 watt might be typical. One section of the relay completes the cathode circuit of V4 and the other sections perform the various functions such as receiver muting, antenna changeover, etc.

Construction

Referring to the front photo of the filter exciter, and the left side of the chassis, the 50 kc bfo can is seen, with the 6BA6 oscillator tube to the right. Directly behind the can, and hidden in the photograph is the carrier level control, R17. In the extreme left rear corner is the 6U8 speech amplifier/phase inverter. To the right of the 6U8 is the 6AR8 sheet-beam balanced modulator and just forward of this tube is the 12AU7 VOX tube. The balance-to-unbalance 50 kc i-f transformer is between the 6AR8 and the 50 kc toroid filter. Partially hidden by the two crystals is the 6BA7 first conversion oscillator. This is the stage that heterodynes

the 50 kc sideband energy to 465 kc. The 465 kc i-f transformers are mounted between the mixer section and the power transformer. You will note that dissimilar transformers were used. Since the Miller #913-C1 has the better selectivity, two of this type are specified on the parts list. Mount T2A back slightly from the position shown so that there will be room between it and the 12AU7 second mixer, for the other transformer.

The power supply components are grouped on the right side of the chassis and include the power transformer, the rectifier tube, filter capacitor, and filter choke. The bleeder/dropping resistors are mounted just forward of the power transformer, vertically, under the chassis.

The second mixer, a 12AU7 is located forward of transformer T2B, next to the balance trimmer and vfo input jack. The 6AG7 output stage occupies the center area of the chassis.

The controls, from left to right, are volume, function (SB-Tone-AM), plate tuning, grid tuning, sideband selector, and power. The clear space at the front of the chassis can be used for mounting a vfo. The input jack is forward of the crystals.

On the underside of the chassis, it will be seen that many of the controls are mounted near their associated circuits, with shaft extentions added for the panel knobs. These controls include the sideband switch, 6AG7 input and output tuning, and the AM-TONE-SIDEBAND switch. Additional front panel controls include the volume control and on-off switch, plus the microphone jack and pilot light. The carrier balance control and VOX gain are mounted on the rear apron. As stated earlier, the carrier insertion level control is mounted on the top of the chassis near the carrier oscillator and balanced modulator. Once this control is set for proper carrier to modulation ratio, it will not require further adjustment.

No problems should be encountered if the layout is followed rather closely, but the use of the section shields in the positions shown in the bottom photograph, is highly recommended. The section running front to back, shields the input and output of the 50 kc filter and the 6AG7. An "outrigger" to the right shields the final from the sensitive 50 kc circuits. An additional plate serves as a mounting for the tuning capacitors and sideband switch and also shields the second mixer from the 400-500 kc oscillator. With this arrangement, the unit is rock solid and no instability can be noted. Components associated with the audio, VOX, and 50 kc oscillator circuit are mounted on terminal boards near the associated tubes, and no shielding was found to be necessary in this section.

Alignment

The alignment is reasonably simple for a unit of this complexity. First, let's establish that the audio is working correctly. Connect the oscilloscope from the junction of C6 and R8 to ground. Place the function switch in the AM or SIDEBAND position and touch the mike jack. A distorted 60 cycle waveform should be noted on the screen. Flip the switch to the TONE position and check for a pure sine wave, of approximately 1,000 cycles. Then, connect the scope to pin 1 and 2

of the balanced modulator (6AR8-V2) and observe an approximately equal sine wave on each electrode. The phase (if you care to check it) will be "180° out".

When the audio is checked out and working properly, connect the oscilloscope to pin 7 of V3, the first mixer. Remove the crystals from the oscillator circuit, and turn the volume control to off. Set the carrier insertion control (R17) and the carrier balance control to full counter-clockwise. A low amplitude r-f pattern should be seen on the screen. Adjust the slug in T4 (carrier oscillator coil) for maximum with the slug almost all the way into the coil. Adjust the two slugs on T1 for maximum (the slugs will be almost all the way out of the coil).

The 50 kc oscillator frequency should be set with an accurate oscillator, using the Lissajous method. However, if none is available, the following method will be satisfactory. Set the carrier balance control (R16) for a null in carrier output with the function switch on SSB. Then switch to TONE and bring up the audio gain until

The r-f output jack is in the center of the chassis, with the carrier balance and VOX gain to the right. The carrier insertion control is between the two tubes at the right side of the chassis.

a pattern (similar to a DSB waveform) is noted. Then adjust the slug on T4 in a *counter-clockwise* direction for minimum ripples on the trace, consistent with reasonable amplitude. This will be approximately the 20 db point on the filter skirt, and about 50 kc. Bring up the carrier insertion and audio gain control for a typical AM wave form. Retain this setting.

Next, move the scope to the common cathode circuit of V4 (pins 3 and 8) and replace the crystals in the oscillator circuit. Adjust T2A and T2B for maximum amplitude of the 465 kc. signal. This waveform should be a replica of that seen in the previous step. It should be noted that the oscilloscope must have good performance in the 500 kc region since the signal is not being demodulated. Any of the so-called "Color TV" scopes should be fine. Turn off the modulation and check for leakage from the crystal oscillator. The r-f should be at least 100 times down from the main signal.

To complete the alignment, connect a source of r-f energy between 3.335 and 3.535 kc to the vfo input jack (J3). The level should be around 10 volts peak. Load the output of J2 with a 50 ohm carbon resistor. If the oscilloscope does not have suitable response at 4 mc it will be necessary to use a "Q" Probe or other r-f pickup device. An oscilloscope can be connected directly across the load resistor, or the "Q" Probe can be coupled to the output tank coil.

Tune the input and output capacitors associated with V5 for maximum amplitude of the waveform observed earlier. You will no doubt note "crud in the picture" indicating presence of the vfo signal. Remove the crystal again, and note that the signal remaining comes from the vfo. Adjust capacitor C19 for minimum r-f on the screen. A complete null should be possible. Replace the crystals and readjust all tuned circuits (except T4) for maximum amplitude. Repeat the T4 adjustment procedure with the function switch on SSB since the oscilloscope loaded the filter somewhat in the first adjustment. Then set the carrier insertion control and audio

gain control for the best modulation, as observed on the oscilloscope. This completes the adjustment of the toroid filter exciter.

A converter suitable for heterodyning the output of this exciter, to the higher frequency Amateur bands was described in the book Single Sideband Techniques, First Edition.

3.4 CRYSTAL FILTERS

The advent of crystal filters for Amateur use made possible the generation of single sideband signals in the region of 400 to 500 kc, and eliminated the necessity of the extra heterodyning operation. Inexpensive crystal filters constructed from the war surplus FT-241 type low frequency crystals are very popular with the "do it yourself" hams. These CT cut crystals have been plentiful and relatively cheap for a number of years and are in the hands of many Amateurs. The general run of the mill war surplus crystal may or may not be good. Experience has shown that approximately one out of four of these crystals are usually defective in one way or another. When ordering crystals from a "surplus house" it would be prudent to order several extra, just in case.

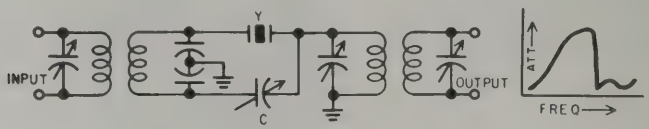


Fig. 3.4A—The crystal arrangement as used in most modern communications receivers.

The most common filter circuit in use is shown in fig. 3.4-A. This is a single crystal arrangement that is standard equipment in most communications receivers. The response curve for a typical setting of the phasing capacitor C is shown at the side of the filter schematic. Varying C will change the shape of the main pass band lobe considerably, but the frequency of the main response peak will remain fixed. However, the sharp null or notch will be moved as the value of C is changed. When C is equal exactly to the shunt crystal holder capacity, the null will not be present in the response curve shown. At this setting of C, the passband will have its sharpest characteristic. As C is varied either above or below this balancing value, the notch will appear either on one side of the center frequency or the other. As C is varied still further from its balanced setting, the notch will get closer to the center frequency of the filter response. The notch cannot come any closer to the center frequency than about 200- 500 cycles. This accounts for the often noticed inability to pull out heterodynes that are close to the desired signal frequency. The notch of the filter, while quite deep, is sometimes broad across the top of the "slot". The variation in pass band shape is also annoying at times. At best, this circuit is not suitable for use in sideband transmitter to eliminate the unwanted sideband.

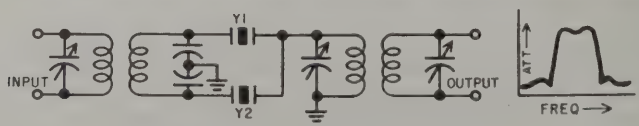


Fig. 3.3B—The popular half lattice crystal circuit.

The *half lattice* filter starts to approach an ideal type of characteristic as shown in fig. 3.4-B. The half lattice is a derivation of the full lattice shown in fig. 3.4-C and theoretically both have the same response. In the half lattice,

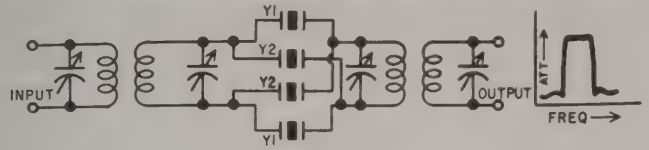


Fig. 3.4C—A full lattice crystal arrangement.

the frequencies selected for the crystals Y1 and Y2 must be 2 to 3 kc apart. The exact difference is not important, but good active crystals should be used. Once aligned, the filter will have a band width approximately 1.2 times the separation of Y1 and Y2. The sharp notches either side of the passband are quite deep (about 60 db for a single section filter). The filter should be aligned so that the drop off outside the passband is sharp while, at the same time, the extra response lobes beyond the deep notches are kept as far down as possible. The actual alignment usually ends up as a compromise between steepness of skirt response and low outer lobe response. These two conditions generally do not occur simultaneously. The correct impedance for the two crystals to "look into" at the primary of the output transformer is just half of that presented by the secondary of the input transformer. Thus, it is often advantageous to place a resistor across the terminals of the primary of the output transformer in each section of the half lattice filter. The actual value must be determined by experiment and will be in the order of a few hundred thousand ohms. The resistor will tend to give a better over all response curve shape. Two sections of a half lattice filter are usually necessary to obtain enough selectivity for SSB generation. This type of filter can also be used to advantage in the 455 kc i-f amplifier of communications receivers. The additional selectivity obtained with this simple crystal filter will really help the old clunker separate stations. An operational circuit for a modified half lattice filter that can be used in receivers or exciters is shown in fig. 3.4-D.

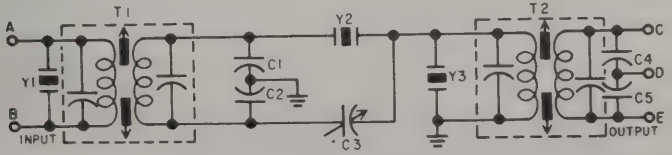


Fig. 3.4D—A modified half lattice filter. This is the heart of the W1JEO exciter described in the ARRL publications.

Parts List for the modified half lattice filter.

C1, C2, C4, C5—100 mmfd. silver mica
C3—3-30 mmfd. compression padder
T1, T2—455 kc slug tuned i-f transformer (Miller 12-C1)

To pass lower sideband: Y1—452.8 kc (Channel 326), Y2—448.6 kc (Channel 323), Y3—450.0 kc (Channel 324).

To pass upper sideband: Y1—447.2 kc (Channel 322), Y2—451.4 kc (Channel 325), Y3—450.0 kc (Channel 324)

The circuit in fig. 3.4-C is a full lattice filter. The same comments apply to this filter as applied to the half lattice filter with the exception that the crystal labeled Y1 must be as close to the same frequency as possible and the same is true for the Y2 crystals. These crystals should be matched to within at least 50 cycles, closer if possible. The frequency difference should be 2 to 3 kc, as in the half lattice. Two tandem connected filter sections will give approximately 60 db attenuation of all signals outside the passband.

The Modified Half Lattice

Fig. 3.4-D illustrates a very popular form of crystal filter. The basic device is the half lattice shown in fig. 3.4-B but with the addition of two crystals connected across the i-f transformer windings. The extra crystals are selected so that their series resonant frequencies fall in the middle of the outer lobes that are undesirable in the basic filter characteristics. The extra crystals need not be restricted to two in number, but as many as six or eight may be used if necessary.

The modified half lattice of fig. 3.4-D is the same as the filter used in the now classic three tube exciter designed by Edmond F. Edmunds, W1JEO. We all owe Edmunds a debt of gratitude, for his W1JEO exciter dramatically showed how simple the generation of a single sideband signal could be, and prompted many Amateurs to radiate an SSB signal during early stages of the art. Anyone contemplating the construction of a crystal filter exciter might do well to construct this unit. The three tube and four crystal W1JEO exciter did have several disadvantages, as you can well imagine. The alignment was a compromise one, in which best sideband suppression did not occur when best pass band characteristics were realized. The slope of the filter characteristics would not be steep enough when the adjustments for maximum undesired sideband suppression were made. Generally, there would also be a large dip in the middle of the transmitted sideband characteristic. The filter circuit of 3.4-D (and the W1JEO exciter) can be successfully modified as follows: Referring to Edmunds' original alignment technique as described in the ARRL publications, carefully adjust for best pass band shape without worrying too much about the suppressed sideband portion of the curve. Then, note the frequencies where the humps (or "pop-up") occur in that part of the curve. Now, additional FT-241 crystals are chosen so that their series resonant frequency is located at the point where additional attenuation is needed. The extra crystals are connected in parallel with the input crystal (Y1 in Fig. 3.4-D), the output crystal (Y3) or in both places. Ken Stiles, W2MTJ, originally conceived these modifications and suggests that additional crystals may be used to improve the steepness of the filter cut-off characteristic next to the carrier frequency.

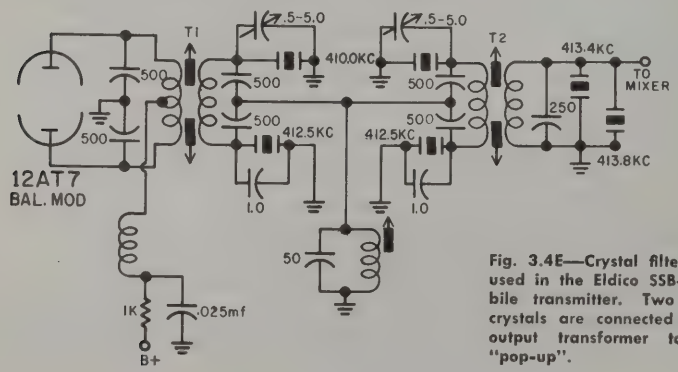
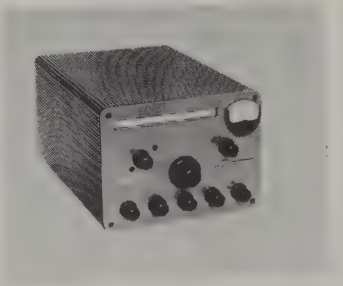


Fig. 3.4E—Crystal filter circuit as used in the Eldico SSB-100M mobile transmitter. Two additional crystals are connected across the output transformer to suppress "pop-up".

Additional sideband suppression may be obtained in the following manner, also; Select a crystal between the carrier frequency (Y3) and the rejection frequency (Y1). Then, choose another crystal that is very close to the carrier suppression crystal (Y3) and connect these crystals in shunt with the transformer windings. As in the original design, Y2 is the pass band crystal and is used to pass the desired frequencies. For upper sideband generation, the crystal is one channel higher than the carrier crystal (Y3). Crystal Y1 is the lower sideband suppression crystal and is two channels below the carrier frequency crystal.

An application for quartz crystals in a sideband filter is shown in fig. 3.4-E. This filter is incorporated in the Eldico SSB-100M mobile transmitter. It is a modified lattice with coupling between the transformer center-taps. The degree of coupling is controlled by a slug tuned inductance, shunting this path. Two additional crystals are connected across the secondary of the output transformer to reduce the "pop-up" mentioned earlier.



The Eldico SSB-100M mobile transmitter emloys a crystal filter centered on 410 kc for suppression of the unwanted sideband.

Determining Crystal Frequencies

The series resonant frequency of an FT-241 low frequency crystal may be determined in the following manner: Connect the crystal in series with the hot lead of a BC-221 or LM frequency meter and the vertical input of an oscilloscope as illustrated in fig. 3.4-F. A sensitive r-f vacuum tube voltmeter or other indicating device that will respond to the 400-500 kc frequency may also be used. Slowly tune the signal generator through the range of the crystal until a sharp kick is noted on the oscilloscope or VTVM. This sudden rise indicates the series resonant frequency of the crystal. Carefully check the frequency of this peak. Shunt circuit capacity should not shift the frequency, so no great care need be taken in arranging the leads other than the usual precautions of keeping them short.

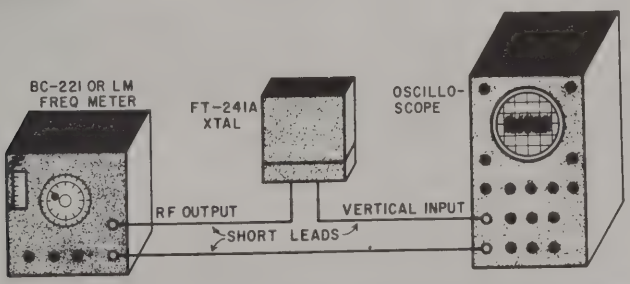


Fig. 3.4F—Test set-up for measuring the series resonant frequency of surplus FT241 crystals.

War surplus crystals have been described, because of their low cost and availability. A chart showing the normal frequency of the two digit (black) and three digit (brown) crystals is shown in fig. 3.4-G. The three digit crystals are the most popular since they provide nearly optimum bandwidth for sideband filters. The equipment that utilized these crystals (in a phase modulated oscillator) multiplied the basic frequency by 72 times to provide enough "swing" for frequency modulation. Therefore, the frequency marked on the case will tell the fundamental frequency simply by dividing by 72. The basic frequency runs from 375 to 525 kc in steps of 1.388 kc, which corresponds to an output frequency of 27.0 to 37.8 mc. The two digit channel numbered crystals have fundamental frequencies which are approximately 1.9 kc apart and are multiplied 54 times. This separation is too great for the standard three crystal Edmunds' filter. They are very useful, however, when used in conjunction with the three digit series, to fill in the curve, provide additional attenuation, or bring about a steeper skirt response.

CHANNEL NUMBER	FREQUENCY IN KILOCYCLES
320	444.4
321	445.8
322	447.2
323	448.6
324	450.0
325	451.4
326	452.8
327	454.2
328	455.6
40	444.4
41	446.3
42	448.1
43	450.0
44	451.9
45	453.7
46	455.6
47	457.4

Fig. 3.4G — Frequencies for common FT-241/A crystals near 455 kc. The three digit types are colored tan while the two digit types are black.

Changing Crystal Frequency

Even with the assortment of "rocks" available from the war surplus stocks, it is sometimes necessary to shift the frequency. A close inspection of the type FT-241 crystals will disclose that conventional contact plates are not used. In their place, silver is plated on either side of the crystal blank and a fine wire soldered at right angles to the plating. Because the flat faces of the crystal are not exposed for grindings, as is usually done, increasing the frequency may be accomplished by lightly grinding the edges. This must be done evenly on both edges, and at the same time maintaining the squareness of the blank. It is recommended that the support wires be carefully unsoldered from the pins at the bakelite bases for such grinding operation. Experience has shown that the FT-241 crystals are quite easily damaged by movement of the short wire leads which are soldered directly to the silver plating on the crystal blank. It is interesting to note that these leads are factory soldered in an accurate positioning jig by a blast of hot air. Therefore, when grinding the crystal, due care should be exercised to avoid breaking these critical connections. It is also possible to etch the crystals to a higher frequency.

A suitable method was described by Robert B. Kuehn, WQHKF, in the November, 1957 issue of CQ Magazine.

3.5 HIGH FREQUENCY CRYSTAL FILTERS

Crystal filters for the standard i-f (455 kc) are quite popular because of the availability of war surplus crystals. It is entirely possible, however, to construct filters for higher frequencies. Work along these lines has appeared in the VK/ZL publications, but it was not until the introduction of the Hallicrafters HT-32 that such filters received much notice in the States. The HT-32 uses a crystal lattice filter system with a design center of 5 mc. The fact that such filters are practical is very evident from listening to the quality and suppression of the HT-32.

Roll-your-own

War surplus stores not only stock the low frequency crystals but usually have bins of higher frequency crystals, too. Frequencies between 8.5 and 9.0 mc are plentiful and can be obtained for next-to-nothing since they do not multiply into any ham band, and are considered "dogs". Pick a frequency near 9 mc and purchase a dozen or so "rocks". You will find that the "spread" is such that a simple filter can be made up without grinding. Pick two crystals about 2 kc apart and try them to the half-lattice circuit shown in fig. 3.4-B. This will produce a characteristic curve with a shape factor of approximately 4.1. The 6 db bandwidth will be about 3 kc and 12 kc wide at 60 db down. Suitable coupling transformers can be made by retuning or padding the Miller #1463 10.7 mc interstage transformer.

A better shape factor will be obtained with the full lattice configuration but this requires knowing the crystal frequencies quite accurately. The calibration technique described in fig. 3.4-F will not be suitable because the frequency meter harmonics are quite weak at this frequency. The crystal can be connected in series with an accurate communications receiver antenna terminal. At the resonant frequency of the crystal, the "S" meter will read considerably higher than off resonance. Another technique is to construct an oscillator and measure the output frequency. This frequency will not be the series resonant frequency exactly, but will show the relationship of the crystals to each other. Thus, if two crystals oscillate 2 kc apart, it is quite likely that the series resonant frequency of the two crystals is 2 kc apart. The exact frequencies are unimportant. Naturally, when comparing crystals, the oscillator circuit should be left alone so that each crystal will be subjected to the same circuit conditions.

Commercial Filters

At this writing, there are three companies known to be actively engaged in manufacturing 9 mc crystal filters. The price of these units range between $25 and $50, depending on the complexity and number of elements in the filter.

Blackhawk Engineering Co., Box 146, Janesville, Wisconsin — 9 mc, 4 kc bandwidth.

Hycon Eastern, 75 Cambridge Parkway, Cambridge 42, Mass.— Model 9 MA, 6 db — 3 kc, 60 db, 12 kc.

McCoy Electronics Company, Mount Holly Springs, Penna.— Shape factor 2.0 9 mc—3 kc—6 db, 6 kc—60 db.

Because of low production costs for high frequency crystal filters you can expect to be seeing them in new sideband equipment and even kits.

A High Frequency Filter Exciter

The advantages of using a 9 mc filter for suppression of the unwanted sideband should be obvious. The 9 mc center frequency allows the use of a 5 mc vfo to heterodyne the sideband energy directly to the 4.0 or 14.0 mc band, depending on the tuning of the output circuit. The use of a high frequency minimizes the possibility of spurious beats that are created in each conversion.

Fig. 3.5-A is an operational schematic for a 9 mc filter exciter. It should be stressed that the circuit is operational, that is, it should work as shown, but has not been wired and tested by the author. The microphone voltage is amplified in a two stage R-C-coupled stage, using a 12AX7, and applied to a cathode coupled balanced modulator. The carrier is generated in a crystal oscillator using a ½ 12AT7. Sideband switching is accomplished by placing the carrier on either side of the filter passband. L2 resonates the oscillator and energy is coupled to the second half of the 12AT7 which acts as a cathode follower. A potentiometer in the cathode circuit allows carrier insertion for tune-up. The oscillator energy is also applied to pin 7 of V2. The oscillator signal, inverted in phase, appears at pin 6 of V2, and is highly amplified. The signal is also coupled to the 1-2-3 section of the same tube through the common cathode resistor. This energy appears at pin 1, greatly amplified, but 180° out-of-phase with pin 6. Adjustment of the trimmer capacitor will allow carrier cancellation. If complete carrier cancellation cannot be obtained, a small carbon resistor inserted in series with pin 6 of V2 will lower the signal amplitude without creating unwanted phase shift.

The filter coupling circuit shown will have to be modified to suit the individual filter. Some units will require resonating the input and output, while others are pretuned. The input to the filter is a DSB signal which is effectively converted to a single sideband by the crystal network. The signal is again applied to a mixer and is converted to the operating frequency by heterodyning with the 5 mc. vfo. The operation of V3 is identical to V2 but complete cancellation of the vfo energy will not be possible. This spurious signal is eliminated in the selective tuned circuits of the output stage. A 15 mc trap is included to suppress the third harmonic of the vfo. The circuitry associated with the 6AG7 output stage was "stolen" from the Central Electronics 10A. Coils L3 and L4 are resonated to the operating frequency with the 365 mmfd. variable capacitors.

3.6 MECHANICAL FILTERS

The Collins F455*-31 series of mechanical filters may be used to provide rejection of the unwanted sideband. This unit is an electro-mechanical bandpass filter consisting of an input transducer, a resonant mechanical section comprised of a number of metal discs, and an output transducer. The input and output transducers serve as electrical-to-mechanical and mechanical-to-electrical coupling devices and when resonated to the operating frequency, do not affect the selectivity of the tiny metal discs. An electrical signal applied to the input terminals is converted into a mechanical vibration at the input transducer by means of magnetostriction. The driving coil either aids or opposes the force of the bias magnet in much the same manner as the voice coil in a permanent magnet speaker. The mechanical vibration travels through the resonant mechanical section to the output transducer, where it is converted (magnetostriction, again) to an electrical signal which appears at the output terminals of the filter. A cut-away drawing of a typical mechanical filter can be found in fig. 3.6-A. This filter is of the "bathtub" style of construction and the terminal headers pass through the bottom of the case. These units are hermetically sealed to prevent deterioration of the filter due to oxidation. The bandpass characteristic curve for the new miniature 3.1 kc mechanical filter (Model F455Y-3.1)

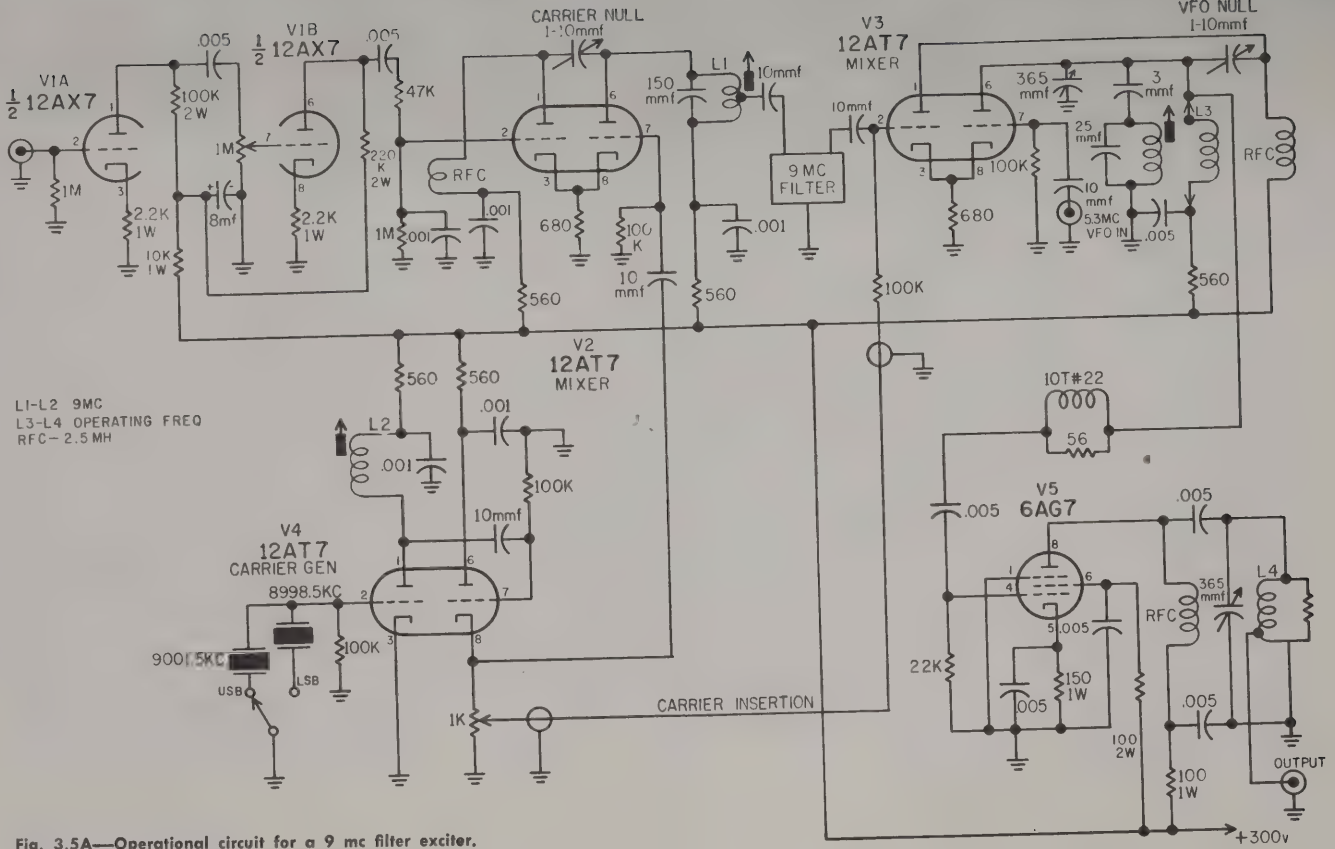


Fig. 3.5A—Operational circuit for a 9 mc filter exciter.

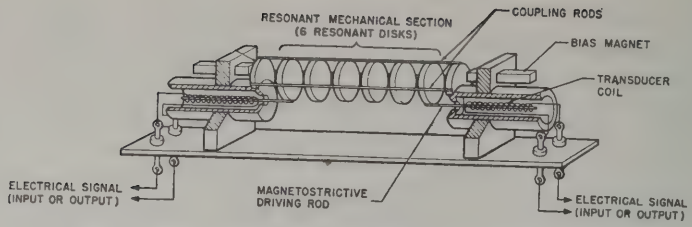


Fig. 3.6A—Internal construction of the Collins "bathtub" style of mechanical filter. Each end contains a transducer coil and bias magnet. Either end may be used for the source or load.

is detailed in fig. 3.6-B. The *shape factor* of this filter (ratio of bandwidth at 6 db to bandwidth at 60 db) is less than 2.0 and is considered extremely good. The frequency of the carrier oscillator is placed at one of the "20 db points" on the filter skirt. The frequencies for these points are approximately 453.2 and 456.9 kc. By moving the oscillator from one frequency to the other, either one of the two sidebands may be placed outside of the filter passband and rejected.

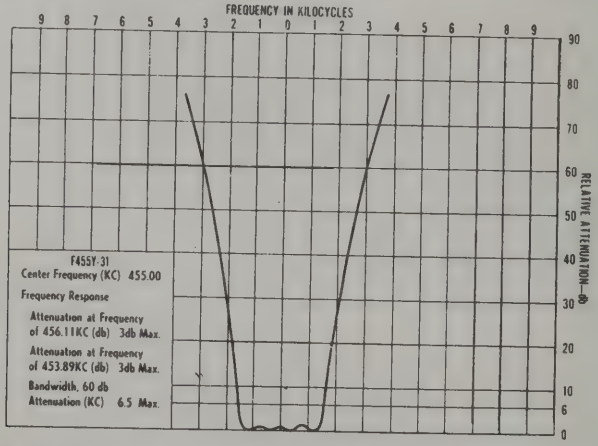
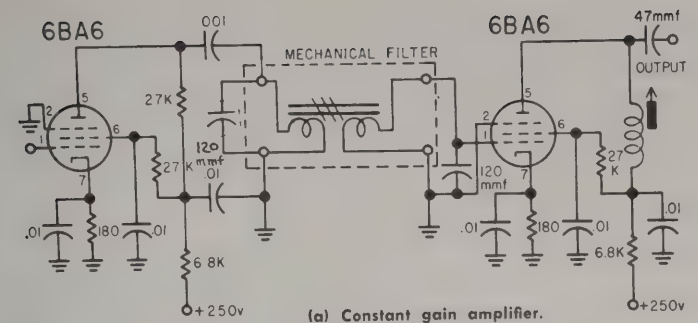


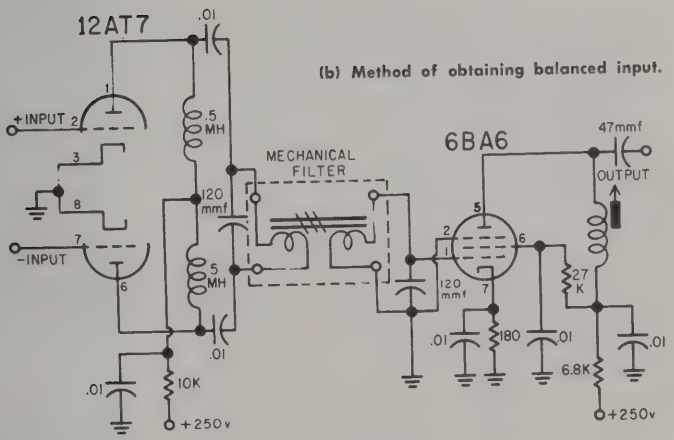
Fig. 3.6B—Characteristic curve for the Collins miniature mechanical filter, Model F455Y-31. The bandwidth at 60 db is 6 kc, and at 6 db is 3.1 kc resulting in a shape factor of less than 2.0.

Mechanical Filter Circuits

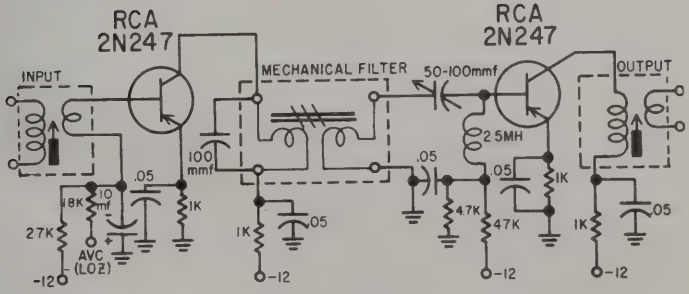
The mechanical filter may be used to separate the sidebands in a transmitter or receiver. It is also well suited for use with transistors in miniature radio equipment. Fig. 3.6-C shows the operational circuits for these applications. The mechanical filter may be used as the coupling element in a 455 kc i-f amplifier as in 3.6-C(a) or to couple the balance modulator to an amplifier stage (b). The balanced circuitry and r-f chokes tend to artificially "center-tap" the primary transducer coil, producing 180° phase difference at each end of the coil. Carrier cancellation, therefore, occurs in the primary coil, while one sideband is transmitted through the filter disc sections. An interesting application for the filter is in transistorized receivers or exciters (c). Although the company recommends that no dc be applied to the coil, it is permissible to connect it to a 12 volt source, providing the current does not exceed a mil.



(a) Constant gain amplifier.



(b) Method of obtaining balanced input.



(c Transistor constant gain amplifier.

Fig. 3.6C—Application circuits for the mechanical filter.

or two. As in other circuits, the primary must be resonated at 455 kc with (in this circuit) a 100 mmfd. capacitor. Because of the very low input impedance of the driven transistor, it is necessary to series tune the output transducer coil. Resonance will occur with approximately 100 mmfd. of series capacity. It should be noted that the value of resonating capacity will depend on the type of filter used. The values shown here are suitable for the miniature 455Y-31 filter.

3.7 A MECHANICAL FILTER EXCITER

A 12AX7, 6AR8 and 6BA6 form the framework of a simple filter type exciter as shown in fig. 3.7-A. A maximum audio signal of 0.10 volts, rms is required at the input grid of the 12AX7 audio amplifier for 5 watts of peak SSB output on 80 meters. The 12AX7 tube serves as a voltage amplifier and phase inverter, while the 6AR8 acts as the balanced modulator. The beating oscillator is a 6AU6 whose frequency is variable over the range of 450 to 460 kc. A double sideband suppressed carrier is generated by the 6AR8 tube, and one sideband is rejected by the Collins mechanical filter which follows the modulator stage. The output of the filter is at a frequency of 455 kc. A 12AT7 double triode mixer stage follows the filter, beating the SSB signal against a 3355 kc crystal conversion oscillator. The resulting SSB signal is on 3810 kc. The SSB signal is amplified by a 12AU7 connected as a cascode amplifier, driving in turn a parallel connected pair of 6AG7's operating class AB_1. The tubes deliver a SSB signal of about 5 watts peak power with negligible high order distortion products.

An audio signal of 0.10 volts rms (easily obtained from the usual variety of crystal microphone) is amplified in the two 12AX7 stages, producing a peak audio voltage of 22 volts to ground between each deflection plate of the 6AR8 (measured at pins 1 & 2). This is sufficient for complete electron beam swing with a plate and accelerator potential of 300 volts. Approximately 5 volts of negative bias is applied to the 6AR8 resulting in a quiescent plate current of 6 ma. per section and a deflector current of less than 1 ma.

A peak-to-peak r-f voltage of 3 volts is required from the 6BA6 electron coupled oscillator. The circuit shown is a modification of the standard Clapp arrangement, wherein the cathode r-f choke is replaced by a 1600 ohm composition resistor. This eliminates a bulky r-f choke, and at the same time provides better waveform from the oscillator. Poor waveform in the low frequency beating oscillator can degrade the performance of the balanced modulator to a great degree. The output from the oscillator is taken from the cathode, since only a small mixing voltage is required.

Sufficient isolation must be provided in the power leads to prevent signal leakage between various stages. Since both the 6BA6 low frequency oscillator and the 6AR8 modulator are "hot cathode" circuits, it is imperative that the filament line supplying these tubes be close to ground potential at the carrier frequency of 455 kc. A .01

Parts List for Fig. 3.7A

C1—100 μμfd variable ceramic padder Centralab 823-DN
C2—30 μμfd variable. Bud LC-1642
C3-C4—.002 μfd silver mica
C5-C6—250 μμfd mica. Series tuning capacitors for mechanical filter
C7—130 μμfd mica. Parallel tuning capacitor for mechanical filter
C8—Output tuning capacitor. 150 μμf Bud MC-1856
L1—2½ mh r.f. choke. National R-100
L2, L3—National XR-50 form, 22 turns #18e. ½" diam, ¾" long
L4—Bud 40-JEL (for 80 meter operation)
RFC-1 — 4 microhenry. J. W. Miller 5221
RFC-2, RFC-3—2½" millihenry. National R-100
PC—50 ohm, 1 watt resistor, wound with 10 turns #22e. wire
MF—Collins 455 Kc mechanical filter, type 455-D31
R1—Audio level control, 0.5 meg
R2—Modulator balance control, 3000 ohms
All .01 condensers are centralab ceramic type DD

Fig. 3.7A—Schematic diagram of the mechanical filter type of exciter.

mfd. ceramic capacitor is placed across the filament pins of the 6AR8 socket and a 0.1 mfd. capacitor is placed across the filament leads of the 6BA6 oscillator. In addition a small r-f choke is placed in the filament line between these two tubes and the high frequency portion of the exciter. As a final step, the B plus line is bypassed with a 0.5 mfd. capacitor, and is decoupled from the rest of the exciter with a 1,000 ohm decoupling resistor.

Low Frequency Oscillator

The 6BA6 oscillator is variable over a range of about 10 kilocycles at 455 kc. This allows the operator to change sidebands by moving the oscillator frequency from one side of the pass band of the mechanical filter to the opposite side. It is possible to transmit an amplitude modulated signal consisting of the carrier and either the upper or lower sideband by placing the carrier at one edge of the filter passband. For normal operation, on SSB, the oscillator is placed on the slope of the passband, close to the 20 db. attenuation points as described earlier.

The "Q" of the inductance used in the oscillator grid circuit determines the value of the silver mica padding capacitors that are connected from grid to cathode of the 6BA6 oscillator tube. With a low "Q" coil, these capacitors may have to be decreased from the indicated value to .001 mfd. apiece to maintain oscillation of the stage. The effect of these capacitors upon the frequency of oscillation is minor, since the frequency determining elements are placed between the coil and chassis ground. The larger the capacity of the padding capacitors, the better will be the stability of the oscillator. A good quality 2.5 mh r-f choke (National R-100) used in this circuit will allow .002 mfd padding capacitors to be used across the tube elements.

The resonant frequency of the 6BA6 grid circuit is determined by the series padding capacitor. Too little capacity at this point in the circuit will cause the 6BA6 stage to stop oscillating. The remedy for this is to remove a few turns from the oscillator coil and to add sufficient padding capacity to restore oscillation.

Sideband Filter and 80 Meter Stages

The 6AR8 modulator stage is coupled to the mechanical filter which is series tuned to provide proper impedance matching to the modulator. The value of the series tuning capacity will vary with the type and model of filter used. The series capacitor removes the dc plate voltage from the filter coils. A stated earlier, the carrier balance of the 6AR8 stage is set by a potentiometer control in the plate circuit.

The output circuit of the mechanical filter is parallel tuned to 455 kc. A SSB signal of about 0.2 volts peak is obtained at this point. A 12AT7 double triode is employed as a conversion mixer to the 80 meter band. The SSB signal and the 3 mc mixing carrier are applied to the individual grids of the triode sections, and the two signals are mixed in the common plate circuit which is tuned to the sum frequency which falls within the 75 meter phone band. A 12AU7 cascode stage follows the mixer and amplifies the SSB signal to a level suitable to drive the parallel 6AG7 tubes to a peak power of about 5 watts.

The 12AU7 cascode stage provides a voltage gain of about 90 without the need for parasitic suppression and/or stabilizing circuits that are usually needed in high gain pentode circuits. The plate circuit of the cascode stage is high C, which helps to further reject the fundamental signal of the crystal controlled conversion oscillator.

The parallel connected 6AG7 output stage employs screen parasitic suppression and series grid resistors to insure stable operation. The stage is not neutralized and tends to be slightly regenerative, raising the intermodulation figure of the transmitter by a measurable amount. The high order products of the transmitter are about 30 db down from the peak output which compares favorably with the majority

of good amateur SSB signals, and which sets a shining example for some of the mediocre SSB signals that can be heard on the amateur bands from time to time.

Exciter Construction

The SSB exciter is built upon a 8" x 12" x 3" aluminum chassis which is divided into compartments to insure that various signals and r-f voltages stay where they belong. A top and bottom view of the exciter is shown in the accompanying photographs. The 12AT7 speech amplifier, 6AR8 modulator, and 6BA6 low frequency oscillator are in one compartment. The center compartment contains the 9002 conversion oscillator and the 12AT7 80 meter mixer. The remaining compartment is divided in two. The smaller section contains the 12AU7 cascode stage, and the larger half contains the parallel 6AG7 amplifier stage. All filament and plate leads to these last two stages pass through bulkhead type capacitors which are mounted in the shield partitions, insuring that the shields are not "short circuited" by power lead coupling loops. When a bottom plate is placed under the chassis, the stage isolation is of a high order.

It is necessary to shield the 9002 conversion oscillator and the 12AT7 mixer tubes, or there will be undesirable coupling around the mechanical filter, between the low frequency stages and the 80 meter stages. This coupling will tend to mask the excellent skirt selectivity of the filter. A shield over the 80 meter conversion crystal is also needed. A minute amount of stray coupling existed between the conversion crystal and the 6BA6 low frequency oscillator tube, even though they were located six inches apart on the chassis. After the unit had been run for a long time, tube shields were placed on all tubes with the exception of the 6AG7's. This resulted in a measurable improvement in the sideband suppression and skirt selectivity of the filter. When a filter provides some 80 db. of unwanted sideband rejection, as the mechanical filter does, stray leakage around the filter can ruin the passband of the unit. Careful attention to details will reduce this unwanted coupling to a minimum.

Exciter Adjustment

Adjustment of the exciter is relatively simple. The 6BA6 oscillator is so tuned that the 455 kc signal falls at mid-scale on its tuning control, which is labelled "Upper-Lower Sideband Selector". The SSB signal at the output of the mechanical filter may be introduced into the i-f section of a monitor receiver by a short length of coaxial line. The oscillator may then be heard as it is tuned across the passband of the mechanical filter. The 80 meter conversion oscillator may now be monitored for operation, and the low frequency oscillator may be heard on 80 meters as it passes across the filter "window"

The next step is to adjust the balance potentiometer in the plate circuit of the 6AR8 stage for minimum 80 meter carrier at the SSB frequency. The sideband oscillator is then tuned just outside the filter passband, and an audio signal is applied to the exciter. If available, a sine wave of about 2000 cycles should be used, at a level of less than 0.1 volts rms.

The signal heard at the 80 meter SSB frequency should be a steady carrier with little trace of the 2000 cycle tone. Any tone that may be heard is a result of the sideband beating with the residual carrier and with high order distortion products. When received on a superhetrodyne receiver with no beat oscillator, the SSB test signal should sound like a carrier, with perhaps just a trace of light tone modulation.

The output circuit of the mechanical filter should be tuned for maximum SSB signal as should the plate circuit of the 12AT7 mixer. The 12AU7 cascode amplifier is next plugged in, and the plate circuit of this stage adjusted for maximum SSB signal. The same step is followed for the 6AG7 stage. The 6AG7's may be loaded with an automobile headlight lamp which will serve as a dummy load. The bulb will light to about half brilliance under full SSB output. When voice modulation is

applied to the exciter, the signal should be completely unintelligible until the receiver BFO is turned on and properly adjusted. The speech should be clear and crisp, and no "buckshot" should appear in the spectrum where the rejected sideband lies. Care must be taken to make sure that the amplifier stages are not tuned to the frequency of the 9002 conversion oscillator, since these tuned circuits offer the only rejection to this signal.

When modulation is removed from the SSB exciter, the lamp load should go out. Any residual glow is an indication of carrier leakage, or of leakage of the conversion oscillator signal through the amplifier stages.

By pushing the exciter, almost 10 watts of output may be obtained, but the higher order distortion products increase sharply in value and the SSB signal is adorned with undesired products falling in the unwanted sideband, and outside the limits of the wanted sideband. At a conservative level of 4 or 5 watts, the signal is clean and sharp.

Low frequency stages are at the left of the chassis. The 6BA6 is in the foreground, with the 6AR8 balance potentiometer behind it. To the rear are the 6AR8, the audio level control and the 12AX7 speech amplifier. To the right of the mechanical filter are the 80 meter mixer and oscillator stages. The conversion crystal is inside the small can. At the right are the cascode and parallel 6AG7 stages.

The output amplifier and cascode stages are in the left hand compartments. The center compartment contains the mixer and oscillator stages. At the right are the low frequency and audio stages. The mechanical filter (atop the chassis) bridges the right hand compartment shield. The input circuit of the filter falls in the low frequency compartment, while the output circuit falls in the mixer compartment.

3.8 A SSB TRANSISTOR TRANSCEIVER

Transistors have been improved and developed to the point where they are entirely practical for many applications presently using vacuum tubes. For sideband service they a "natural", since the basic philosophy is to do as much work as possible at low level stages, leaving the amplification of the output signal to high power linear amplifiers.

This transceiver uses standard, readily available, and inexpensive parts in all sections of the receiver and exciter. The transistors and transformers are usually available from local distributors, or can readily be obtained by writing the mail order supply houses.

The device is called a transceiver since the high frequency conversion oscillator and BFO are common to the receiver and exciter. Although many of the transistors

could be made to do "double duty"; it was felt that the overall complexity would be less by using separate receiver and exciter r-f circuits, with common oscillators. Type PNP transistors are used exclusively in the receiver portion, while type NPN transistors are used in most of the exciter circuits. There is no particular reason for this distinction; it occurred only because the particular transistors were readily available. It does illustrate that both types are equally useful, however. The low priced RCA 2N247 "drift" transistor shows markedly superior performance as a mixer, and it is recommended that this unit, or the newer 2N371 — with improved high frequency characteristics, be used as a mixer.

For initial experiments, the lowest priced "experimenter" variety of transistors were used. As the experiments progressed into the r-f range, it was evident that these inexpensive types were noticeably inferior to the "converter" types, with cut-off frequencies in the neighborhood of 7-9 mc. Even at 455 kc, although the low priced transistors would oscillate readily, there was a tendency toward drift for several minutes. This effect was not apparent with the better transistors. Many of the less expensive transistors have a high hiss level when used as a microphone amplifier, whereas the hiss was not evident with the more "exotic" types. Since the "converter" types cost only a dollar more than the cheap transistors, and were superior at audio frequencies as well, it was decided that true economy lay in using them for all low level applications. For power applications, such as driver and output stages, it is wise to use transistors designed for this service. The 2N188 output transistors used in the receiver are no larger than the "experimenter" type, but give adequate speaker volume.

Circuitry

The receiver is a conventional superheterodyne circuit as shown in fig. 3.8-A. The BFO is used as the exciter carrier generator, and after modulation and filtering of the unwanted sideband, is again mixed with the receiver high frequency oscillator. If this scheme is not familiar to you, a little thought will make it clear that this insures that the exciter will transmit on the, exact same frequency (and sideband) as the receiver is tuned to.

The antenna is coupled to the mixer through a band pass circuit, which is nothing more than two tuned circuits slightly overcoupled. This insures that unwanted signals (such a broadcast band stations) will not reach the mixer and cause cross modulation. The secondary of the band pass transformer is capacity coupled to the base of the mixer, a type 2N247. A small value of capacity is used to avoid loading the tuned circuit with the low base input impedance. A capacitive voltage divider, or tap on the coil (to match impedances) would be a refinement, but appears to be unnecessary. Injection from the oscillator is obtained by means of a small capacitor from the base of the oscillator. The injection level does not seem to be critical. The mixer output is coupled to the selective i-f amplifier.

The oscillator is particularly useful for frequencies above the audio range. It gives better stability than other circuits that were tried, and is of the convenient two terminal type. A high LC ratio and a small tuning capacitor spreads the 3.8- 4.0 range across 180° of the dial. To minimize circuit loading the output is taken from the emitter of the oscillator. This point also feeds Q3, a type 2N140 transistor, used for isolation. This stage (Q3) is connected in a grounded collector (analogous to a cathode follower) and feeds the high frequency signal to the exciter. The oscillator transistor is operated at a very low level to minimize drift. The stability is equal to most communications receivers. Some drift will be noted in the first few minutes of operation, until the junctions reach their final temperature.

The mixer is coupled to the first i-f stage with a miniature transistor i-f transformer. This stage (Q4, a 2N140) is followed by a conventional half lattice filter section. The input to the filter is a "tube type" miniature i-f transformer, with the

secondary balanced by two 25 mmfd. capacitors. The output transformer of the lattice is another transistor type, in order to produce an impedance step-down to drive the base of the second i-f amplifier. It is interesting to note that it is necessary to have a stage of i-f amplification before the lattice filter. The output level of the mixer is quite low. If the filter is placed directly after this stage, it will attenuate the output of the mixer. The noise level of the following i-f stage will be equal to or greater than the output of the mixer on weak signals. Naturally the noise figure of the receiver would be impared. With the arrangement shown, the signal to noise ratio of the receiver is very good.

The rest of the i-f section consists of cascaded transistor amplifiers. The gain is more than adequate and some reserve is available. Less than optimum bias is used on the first i-f stage in order to hold the gain to an amount which will provide noise output with no signal, but low enough to eliminate over-all system feedback. With full gain, oscillation takes place around the whole i-f system and a more elaborate

Parts List for the transistor SSB transceiver.

C1—50 mmfd.
C2, C13, C14, C40, C41—25 mmfd.
C3—5 mmfd.
C4, C5, C7, C16, C17, C18, C20, C28, C32, C36, C46, C49, C51, C52-C53-C54—.01 mfd.
C6, C8, C19, C23, C38, C39—20 mfd., 15 WVDC
C9—150 mmfd.
C10, C24—50 mmfd. variable
C11, C12, C29, C44, C45, C47, C48, C50—100 mmfd.
C15—2 mfd., 15 WVDC
C21, C22—10 mfd., 15 WVDC
C27, C34—.002 mfd.
C30—.001 mfd.
C31—33 mmfd.
C33 — 250 mmfd. compression padder capacitor
C35, C37—.0015 mfd.
C42, C43—800 mmfd.
Note: all capacitors less than 1 mfd. are disc or tubular ceramic.
CR1-CR2—1N48 germanium diode

F1—¼ ampere, 3AG slo-blo fuse
J1, J4—RCA Phono connector
J2—Miniature phone jack
J3—4 pin Amphenol microphone jack and plug
Q1, Q14—RCA 2N247 (PNP)
Q2, Q3, Q4, Q5, Q6 — RCA 2N140 (PNP)
Q7 — Raytheon CK-722 (PNP)
Q8—GE 2N107 (PNP)
Q9—GE 2N190 (PNP)
Q10, Q11—GE 2N188
Q12, Q13 — Texas Instruments 903 (NPN)
Q15, Q16, Q17—GE 2N169A (NPN)

L1, L2, L3, L4-30 turns #30 cotton covered, scramble wound on ⅜" form (Miller #4407) — 14.8 — 31 microhenries
R1, R16, R41—270K
R2—15K
R3, R24—680 ohms
R4, R33—33K
R5, R6, R13, R18, R34, R37—1K
R7, R35—4.7K
R8—20K potentiometer, linear taper (Centralab "Radiohm", B-22)
R9—470K
R10, R12, R15, R36, R40—100K
R11, R17, R22, R46, R48—10K
R14, R45—330 ohms
R19, R42—470 ohms
R20—3.3K
R21 — 10K potentiometer, audio taper (Centralab "Radiohm", B15)
R23, R30, R44, R53—47K
R25—3.9K
R26, R29, R52, R54 — 100 ohms
R27, R28—10 ohms
R31—22K
R32—2.2K
R38, R39—270 ohms
R40 — 500 ohm potentiometer, linear taper (Centrafab "Radiohm" B-4
R43, R47—180K
R50—240 ohms
R51—120K
S1a, S1b — two pole, single throw switch—part of R8, the rf gain potentiometer (Centralab "Fastatch" KB-2)

T1—3.8—4.0 mc. bandpass transformer (Miller #6203 modified per text)
T2, T4, T5, T6, T7, T10, T12—Transistor 455 kc. i-f transformer (Miller #2031)
T3, T11 — Miniature 455 kc i-f transformer (Miller #12C-10)
T8—Interstage audio transformer (Halldorson GH-110)
T9—Push pull transistor to speaker transformer (Halldorson GH-4)
T13 — Interstage audio transformer (Halldorson GH-10)
Y1 — War surplus FT-241 crystal, 453.7 kc (Channel 45)
Y2 — War surplus FT-241 crystal, 455.6 kc. (Channel 46)
Chassis—
Dial—National MCN

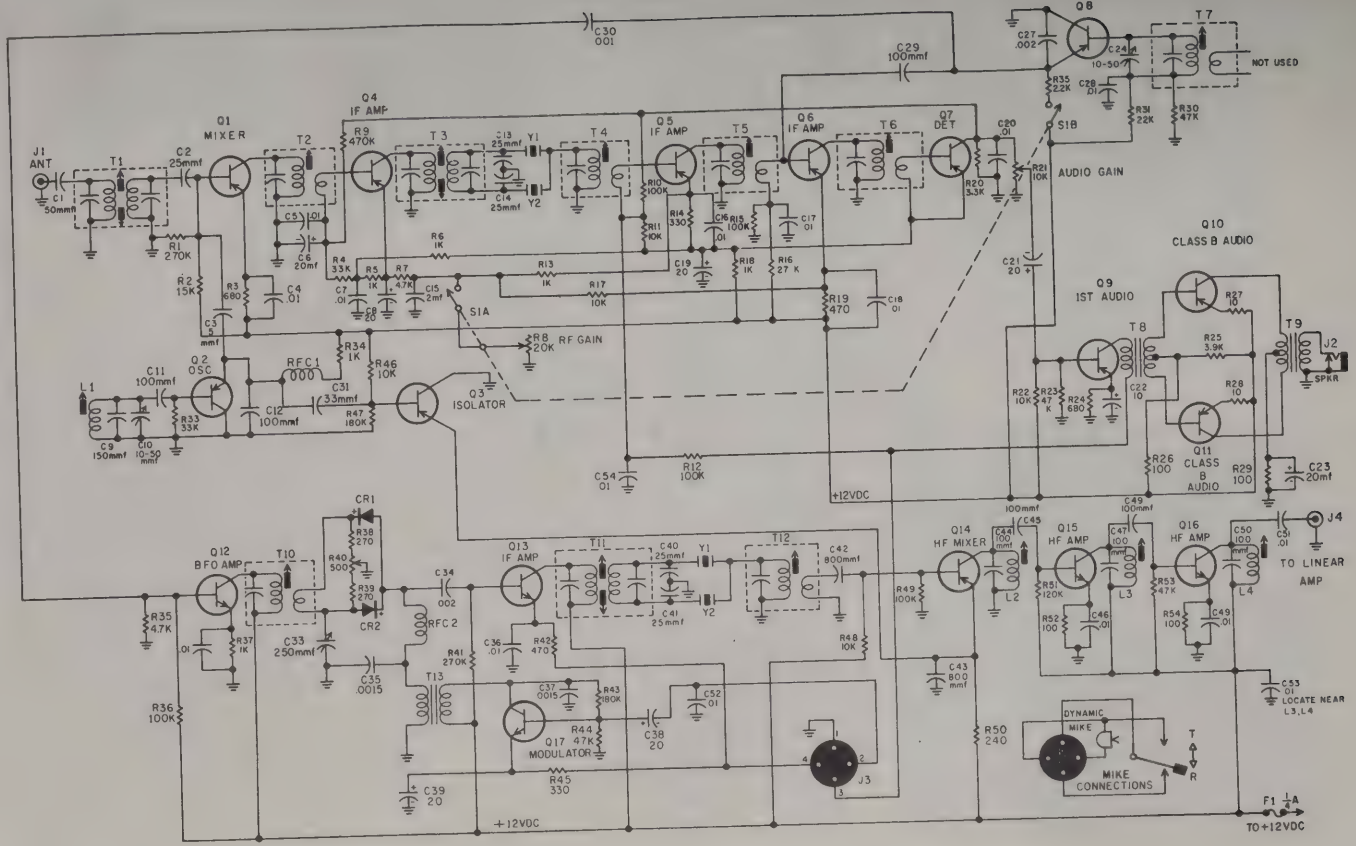
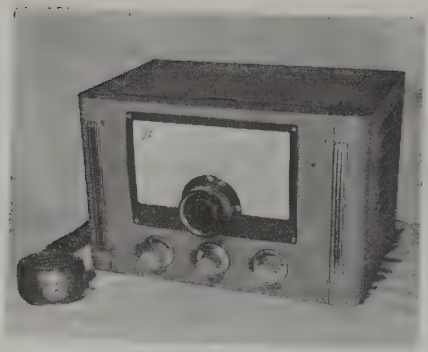


Fig. 3.8A—Schematic diagram of a transistor SSB transceiver.



A completely transistorized SSB transciever for 75 meters. The total power consumption is 42 ma, far less than a pilot lamp! The unit has only four controls, making it very simple to operate. The center knob selects the upper or lower sideband.

decoupling scheme would be necessary. The 2N140 transistors have very low internal feedback capacity and neutralization was found to be unnecessary.

The detector is of the power type. There is no forward bias on the detector transistor and it draws very little current with no signal. Signals delivered by the i-f amplifier drive the base to conduction on the negative half cycle (PNP transistor), detection takes place in the base-emitter junction, and the detected audio is amplified in the collector section of the transistor. The drop across the collector load is proportional to signal strength at the detector. The voltage developed across the collector load is positive with respect to ground and is fed back as AVC voltage to the bases of the first and second i-f stage. Since the base of a PNP transistor must be biased negative for conduction, and the gain varies with the amount of this bias, the positive AVC voltage lowers the negative bias (thus lowering the gain) and effective AVC action is obtained.

The BFO uses the same circuit as the high frequency oscillator. A standard transistor i-f transformer is used as the tuned circuit, and the secondary is unused. A variable capacitor on the front panel tunes the BFO through the i-f filter passband, to allow upper lower sideband transmission and reception. The BFO signal is injected into the last i-f stage (base) in order to provide adequate carrier for SSB demodulation. This also creates additional AVC voltage, but the injection level was adjusted to a compromise which did not lower the effective sensitivity of the receiver, but gave good demodulation.

The audio circuitry is quite conventional, and consists of a volume control at the base of the transistor driver (Q9-2N190). which is transformer coupled to the output audio amplifier. This stage operates class B and the "idling" current is only 10 ma. On voice peaks, this stage will kick up to several times that current.

Exciter Circuitry

The exciter signal begins with the receiver BFO, which acts as a carrier oscillator. This circuit is capacity coupled to a buffer which raises the level very slightly. Its primary function is for isolation between the oscillator and the balanced modulator. The method of balancing out the carrier is about the simplest form and employs only two diodes. The two units were matched by selecting from several, for equal forward and back resistances as shown on an ohmmeter.

A single stage of speech amplification is adequate to raise the mobile dynamic microphone audio to a usable level. The modulation transformer is a transistor interstage unit, with a considerable step down ratio. The transformation ratio does

Alignment

The 455 kc. i-f amplifiers should be aligned first. Connect a signal generator

Top view of the transistor SSB transceiver. Starting at the power plug, and progressing around clockwise, we find the receiver mixer circuit, crystal filter, 455 kc i-f amplifier, with the BFO and 1st audio in the upper left corner. The two transformers are associated with the class B audio amplifier. The exciter portion occupies the right hand side of the chassis, with the audio modulator, balanced modulator, and crystal filter in the upper right hand corner. Just to the right of center is the exciter mixer followed by the linear amplifiers. Directly behind the variable capacitor is the master oscillator and isolation stage.

to the base of Q1 (2N247) and peak T2, T3, T4, T5, and T6 at 455 kc. Then readjust T3 and T4 for the best pass band shape. A few checks with an LM or BC-221 will ascertain when the curve has a flat top and steep sides. Next, tune in the high frequency oscillator on a communications receiver. Adjust the setting of L1 and the number of plates on C10 until the oscillator frequency just covers 3345 to 3545 kc. This will spread 3.8- 4.0 mc over 180 degrees of dial rotation. Adjust the primary of T1 (which has been modified per fig. 3.8-B) to peak at 3950 kc and secondary at 3850 kc. This will produce a bandpass that has flat response in the phone band, but attenuates frequencies outside the band.

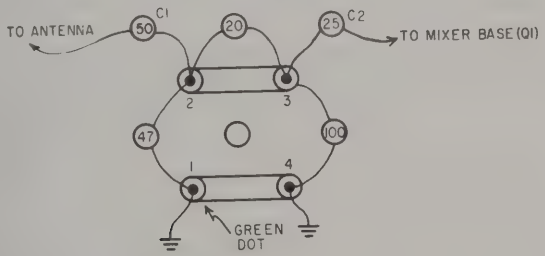
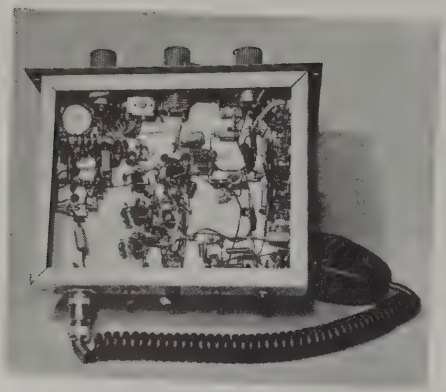


Fig. 3.8B—Modification of the Miller 6203 (4.5 mc interstage) for a bandpass of 3.8 to 4.0 mc. Condensers C1 and C2 are shown in the complete schematic (fig. 3.8A) but the three added capacitors are not.

Turn on the exciter by depressing the mike button. Connect a source of 3.9 mc signal to the base of Q14 (2N247). Peak coils L2, L3, and L4 to this frequency. Then stagger the coils by peaking L2 at 3850, and L4 at 3950. This will produce nearly constant gain over the 75 meter phone band. Next, connect an audio oscillator to pin 2 of the microphone connector and tune in the signal on a communications receiver. Set potentiometer R40 at one end of its rotation. Peak T10, T11, and T12 for maximum signal strength in the receiver. Balance out the carrier by adjusting R40 and C33, alternately, for a null in signal strength (with no modulation). The transceiver is now ready for use.



An underside view shows the circuits to be compact, but not crowded. The unused space near the microphone connected was reserved for the latter addition of a voice operated relay circuit.

not seem to be critical and almost any interstage transformer (designed for use with transistors) can be used.

The crystal filter is driven by the balanced modulator which generates the two sidebands and suppresses the carrier. The crystal filter then removes the unwanted sideband. The filter is the same as in the receiver section and, of course, must be on the same frequency. The usual surplus crystals were used and Channel's 45 and 46 were selected since they were resonant near the i-f frequency. The use of separate filters simplifies the switching problem considerably. Layout of the two filters should be identical to insure that the filter pass-bands will coincide closely. The sideband attenuation measures a conservative 20 db., and the shape of the selectivity curve is quite good. No effort was made to steepen the sides by using additional shunt crystals or tuning the high frequency crystal.

A mixer stage follows the crystal filter. Again the RCA 2N247 (Q14) is employed because of its excellent mixer characteristics. The oscillator is injected into the emitter of Q14 and is obtained from the emitter follower isolation stage (Q3). This configuration is used to prevent frequency modulation of the high frequency oscillator.

The signal frequency amplifiers (Q15-Q16, 2N169's) are driven by the mixer output and will produce about 2.5 volts of drive to the outboard linear amplifier. The "power amplifier", a 2N169A, is coupled to the linear through a short length of coax cable. The linear consists of a 6AK6 driving a power amplifier stage. The capacity of the cable (from the exciter) is tuned out by a resonant coil at the grid of the 6AK6 to extract every possible volt from the exciter. Using this arrangement, there is more than enough drive available to "flat top" the linear amplifier.

Construction

Not much can be said about the chassis layout beyond the photographs and captions. The receiver section occupies the right side of the chassis and the exciter portion will be found at the left (bottom view of chassis). The common oscillator circuits are located in the center of the chassis. On the rear apron of the chassis (left to right) are the microphone connector, exciter output, 12 volt test point, receiver antenna input, and speaker jack. The carrier balance potentiometer is located in the upper left hand corner (bottom view) and the balance capacitor is located near the BFO trimmer. A double pole switch on the rear of the r-f gain control is used to short the AVC circuit and turn on the BFO. A single pole switch on the rear of the volume control turns the 12 volt supply on and off.

Power Supply

In receiving conventional AM, oscillator stability is no problem. In receiving or transmitting SSB, however, the stability of the oscillator is critical. The sad fact is that transistor oscillators are more voltage sensitive than vacuum tube oscillators. When using the super tiny 9 volt battery as a power supply, SSB signals become unreadable when the audio gain control is advanced; voltage fluxuations caused by the current drain of the class B output stage will affect the oscillator and cause frequency modulation. It is necessary to use batteries at least as large as the "penlite" size, with the "flashlight" size preferred. However, it is convenient to use an ac source, as batteries will eventually give out. The ac supply is reliable and flexible. A power plug near the rear apron of the chasis enables either the ac supply, dry cell batteries, or a car battery to be used interchangeably. The ac supply uses a small 25 volt transformer, germanium rectifiers, and a power transistor regulator. This circuit is diagrammed in fig. 3.8-C. A small battery is used as a reference voltage for the regulator and by changing taps on the battery, the regulated voltage may be varied to any desired value.

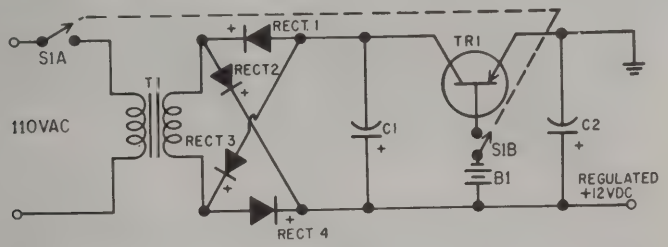


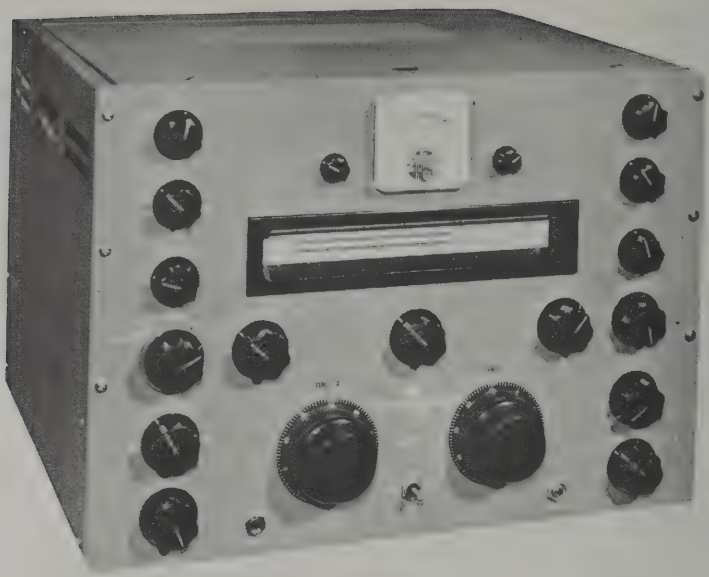
Fig. 3.8C—Regulated power supply for the SSB transistor transceiver.

B1— Three 4 volt mercury cells in series (Mallory TR-233R or equiv.)
C1, C2—50 mfd., 50 volt electrolytic
Rect. 1-4—Germanium rectifiers (General Electric 1N91)
S1a, S1b—DPDT toggle switch
T1—Transformer, 117 vac to 25.2 vac (Stancor P-6469)
TR1—Regulator transistor (CBS 2N256 or RCA 2N-301)

Parts List for the SSB transistor transceiver.

It is obvious that for mobile operation, a regulated supply will be required to eliminate voltage variations under normal driving conditions. The regulator portion of the circuit shown in fig. 3.8-C can be used to regulate the car battery at 10.5 volts by eliminating the transformer and rectifier. This circuit will only work for automobiles with a negative ground.

The equipment was checked out using two sources of power, a 9 volt dry cell battery and a transistor regulated power supply which delivered 13.5 volts. The transceiver works well with either supply. The total drain for the receiver-exciter combination, with the 13.5 volt supply is 43 ma when receiving, and 44 ma when transmitting. This input calculates out to about .6 watts, which is less than the power required to light one pilot lamp! It is obvious that with such a light drain, a dry cell battery is quite practical.



The Cosmophone "35" Bilateral Transceiver. The two large knobs are for tuning the two separate vfo's and the switch near the center of the two controls the function of each vfo. The controls grouped at the right are for the receiver and those on the opposite side are associated with the transmitter. The "35" uses a 6146 linear amplifier output stage.

Conclusions

Operation of the transceiver is suprisingly good and the simplification of controls has been found to be quite convenient. The receiver will reach antenna noise in the daytime, when there are no signals. At night, the r-f control must be cut back for most SSB stations. A considerable amount of listening has been done and this receiver will pick up anything that the regular station receiver can "hear". The "big box" is a double conversion superhet, with a 50 kc i-f strip, product detector, and draws about 150 watts from the power line! Stations report that the signal sounds good, and is normal in every respect. Operation is quite simple — just tune in the station that you wish to talk to, press the mike button, and talk. No need to zero in on the frequency — you're on it!

The author would like to thank Mr. Ed Landefeld, W8DCC, (who designed and constructed the unit) for supplying the schematic and construction data. The excellent photographs taken by Anton Simon, W8DJE, will greatly aid anyone duplicating the SSB Transistor Transceiver.

Additional Notes on Transceivers

The SSB transceiver is gaining wide popularity with Amateurs. Several companies are currently producing these devices, which employ common receiver and transmitter circuits, and many more are expected to jump on the "bandwagon". Some may think the cost of commercial transceivers excessive, but the price becomes slightly less staggering when one considers that if the receiver and transmitter were separate units, the total price would be far greater.

The Collins KWM-1 is an example of a very popular transceiver. Although designed primarily for mobile operation, a power supply is available for operating the device from the ac power line. The circuitry is far too complex to discuss within the scope of this book, but it is worth noting that rather than simply using common oscillators, many of the tubes in the KWM-1 are common to both the receiver and transmitter. The KWM-1 receives or transmits on the same frequency, SSB or CW signals, in the 14.0- 30.0 mc. range. Blocking bias control circuits switch from receive to transmit functions by disabling some of the tubes. Although a mobile device, the circuitry includes vox anti-trip for voice operation.

The Cosmophone "35" is another interesting commercial transceiver. It is the first such device designed expressly for home station use and features all band operation. The "35" has dual VFO's so that independent receiver and transmitter tuning is available; or with the flip of a switch, the transceiver will track on either channel. Fig. 3.8-D is a block diagram of the Cosmophone "35" and shows how many of the circuits are common to both functions. The "35" is a dual conversion superhet, employing a 2.2- 2.8 mc 1st i-f and a 455 kc 2nd i-f amplifier. A mechanical filter eliminates the unwanted sideband and the "Q" Multiplier calls a halt to those annoying heterodynes, when receiving.

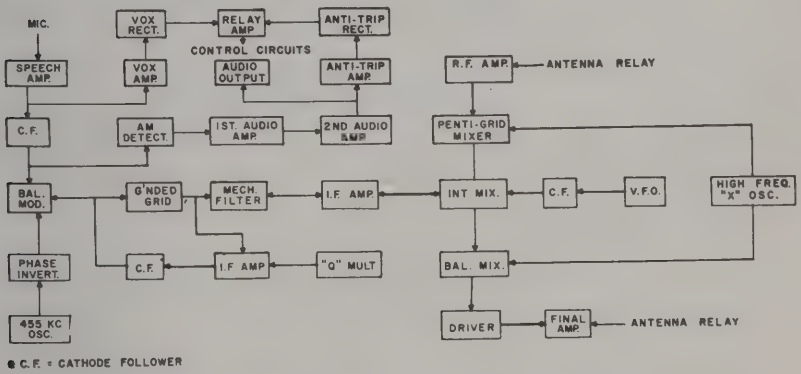
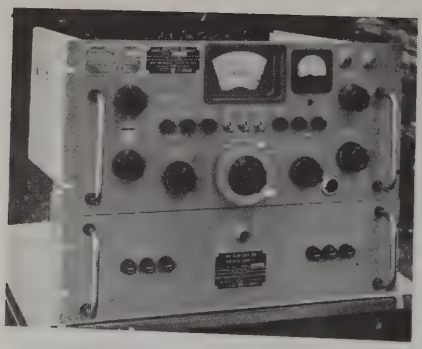


Fig. 3.8D—Block diagram of the Cosmophone "35" Bilateral Transceiver. Note that such circuits as the mixer, i-f amplifiers, balanced modulator and mechanical filter are common to both the receiver and transmitter.

3.9 LINEAR AMPLIFIERS

Regardless of the system used to generate a sideband, a linear amplifier is usually employed to raise the signal to a usable level. One often hears the term "barefoot" on the ham bands, and this means that the operator is using his exciter without a high power linear amplifier. Even the low power output stage of the exciter must be linear.

Great pains are taken to generate a signal that conforms to a certain set of standards. We must faithfully reproduce this signal when boosting it to higher power levels, hence the need for a linear amplifier. Simply stated, linear amplification means that if the input voltage doubles in amplitude (from 1 to 2 volts, as an example) the amplified output voltage must change in the same proportion (20 to 40 volt increase).



Technical Material Corporation sideband transmitter, Model SBE-1. Creation of the sidebands is accomplished at 17 kilocycles through the use of inductive filters.

There are various classes of linear amplifiers, including A, AB_1, AB_2, or B. Ordinarily, Class A amplifiers are used at low levels for voltage amplification of SSB signals. Class AB_1, Class AB_2 or B stages are used for high power final amplifiers. Because of the complexity and importance of the subject, linear amplifiers are discussed, in detail, in a later chapter.

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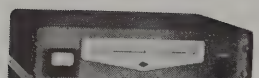
4

The Phasing System

WE HAVE examined, in great detail, the filter method of eliminating the unwanted sideband. Another method, equally popular with Amateurs, is known as the phasing system. Basically, the phasing system is the method whereby the wanted sideband is created, and the unwanted sideband eliminated, by voltage cancellation. An explanation of this system might be more palatable if you recall the old system for eliminating hum pickup in low level equipment. As an example, let's say that a sensitive oscilloscope is displaying a 60 cycle pattern from hum pickup on the test leads. The hum can be completely eliminated by injecting *more* hum into the oscilloscope! Perfect cancellation will occur if three conditions are met: the injected signal must be *exactly* 180° out of phase, it must be on *exactly* the same frequency, and it must be exactly equal in amplitude to the signal picked up on the oscilloscope input leads.

In a phasing exciter, aiding and opposing voltages are used to balance out the unwanted sideband rather than rejecting it, as in the filter system. The voltages must also be of the proper phase, amplitude, and frequency. The phase and amplitude requirements are met by employing phase shift networks, and associated amplitude balance controls. Establishing the correct frequency, in this case, is automatic.



The Gonset GSB-100 exciter uses a combination of phasing and filter techniques. The sideband is generated at 9 mc in the usual manner but a crystal filter is used for additional carrier suppression. The signal is then hetrodyned to within 5 mc of the desired band. A second conversion stage, using a 5 mc vfo heterodynes the sideband to the desired frequency.

4.1 ABOUT PHASING EXCITERS

The phasing exciter employs two, rather than one, balanced modulators with common outputs. Taken separately, each balanced modulator generates a set of sidebands, and suppresses the carrier. In the common output one set of sidebands (upper or lower) will be in-phase, while the opposite sidebands will be out-of-phase. Thus, the in-phase sidebands will "add-up", while the out-of-phase sidebands will cancel. Voltage in the output tank of the *double balanced modulator* consists of one sideband, a cancelled unwanted sideband, and a suppressed carrier. This signal is used to drive succeeding amplifier stages.

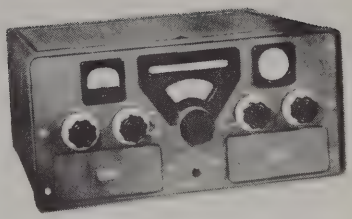
It sounds very simple, and in practice, it actually is. One might wonder why use the filter system at all when it is so easy to generate a sideband in this manner. Although a phasing exciter is relatively simple to construct, and get operating, the adjustments are more likely to drift due to component ageing, and so on. The effect of phase and amplitude errors can be rather disastrous, in the sideband suppression department, as we shall see. This is not considered a serious shortcoming, however, as it is a very simple matter to set the suppression adjustments. The author considers routine maintenance an excellent practice for any exciter, whether filter or phasing.

An advantage of the phasing system, that has not been mentioned, is that sideband suppression can be accomplished at the operating frequency. As we have seen, this is not practical with the filter type exciter. It is possible to feed a source of r-f to an *adapter* which converts the cw energy to a single sideband, suppressed carrier signal. This done right on the operating frequency (yes, even on 10 meters) and the sideband is fed directly to the linear amplifier. No heterodyning process is necessary. An adapter differs from an exciter in that it is designed with an eye towards converting present AM transmitters. The Heath SB-10 is such a device and contains no carrier generators or conversion circuits. In practice, r-f is removed from the transmitter (via a coax cable) at the grid of the final amplifier. The SB-10 converts the energy to single sideband and returns it to the final for amplification. Naturally, the original connection between these two points must be broken.

There are two general philosophies in building phasing type SSB exciters and these are: (1) Build a fixed frequency phasing-type sideband generator and heterodyne the single sideband signal into the Amateur bands of interest and; (2) Build a "fundamental phasing" type of exciter in which the sideband is generated at the same frequency as that desired for the output signal.

4.2 HETERODYNE PHASING EXCITERS

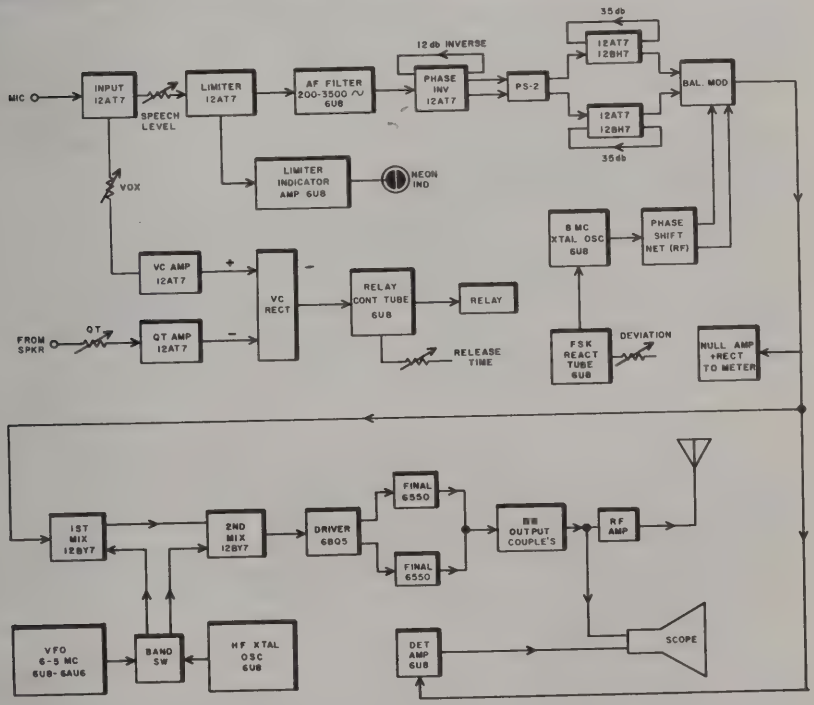
Heterodyne, or frequency conversion stages will be found in phasing exciters, also. To minimize a possible source of phase errors, the sideband may be created



The Central Electronics 100V "Multiphase" exciter-transmitter uses a system of band-pass couplers for one knob tuning. The basic system is the same as that shown in Fig. 4.2A, in that the sideband is generated on a fixed frequency and then heterodyned to the desired Amateur band.

at some convenient frequency and then heterodyned to the desired ham band. This is the system used in the Central Electronics Multiphase, Johnson Pacemaker, and Lakeshore exciters. The large number of these units heard on the air proves that the popularity of the two systems is equal.

All phasing exciters have one feature in common; that is they contain two phase shift networks. Fig. 4.2-A shows the block diagram of a simple phasing exciter. Note that the speech amplifier drives an *audio phase shift network* and the carrier generator delivers energy to an *r-f phase shift network*. In either network, a phase shift of 90° is the "magic number" to bring about sideband suppression in the double balanced modulator. A schematic for a typical double balanced modulator is shown in



Block diagram of the Central Electronics 100V exciter-transmitter.

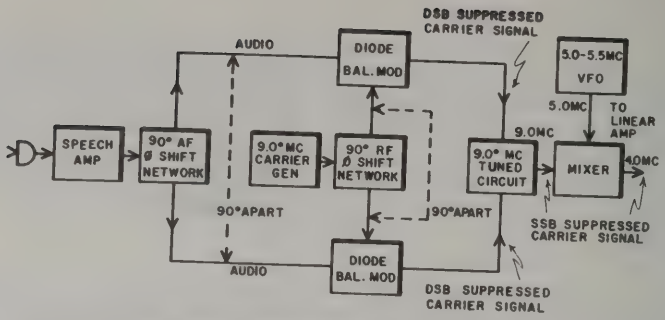


Fig. 4.2A—Block diagram of a fundamental frequency phasing exciter. This is essentially the same system used in the Central Electronics and Lakeshore exciters.

fig. 4.2-B. Audio modulation, shifted 90°, is fed to the modulator diodes through the r-f coupling links and balance potentiometers. Mentally blot out the lower section of the modulator and note that the two upper diodes are tied to opposite ends of the push-pull tank circuit. When the arm of the balance potentiometer is "centered" an equal amount of r-f is fed to each diode, and carrier cancellation occurs in the push-pull tank circuit. When modulation is applied to the diodes, the balance is upset, and the two modulation sidebands appear in the output circuit. Referring to the lower section of the schematic, it will be seen that exactly the same circuit configuration exists. The action in these two sections is identical.

If the audio and r-f signals were in-phase, nothing more than two parallel connected balanced modulators would exist. However, the two coils (L1 and L2) are adjusted in such a manner, so as to create a 90° r-f phase shift between the two potentiometer arms. By phasing the r-f and audio signals as shown, one set of sidebands will add and the other set will cancel. Let's take a specific example: The upper sideband voltages of the two channels will be equal in amplitude, but due to all the phase changes, we have purposely added, the component voltages of the two upper sidebands are exactly 180° out-of-phase. This means that when combined in a linear device like the tuned tank circuit, the upper sideband voltages will cancel each other. However, the lower sideband voltages in both channels will not be out of phase, but will be exactly in phase. These voltages will add up to give a lower sideband voltage twice as large as that existing in either channel.

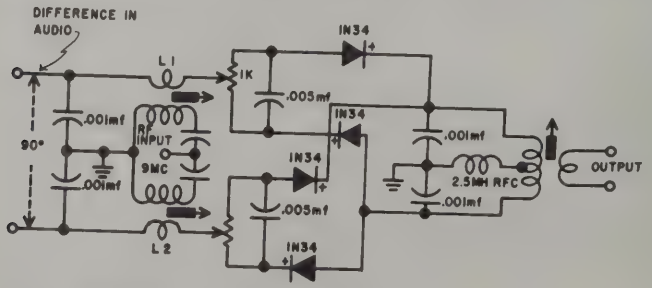


Fig. 4.2B—Balanced modulator portion of a phasing exciter. L1 and L2 are tuned so that a 90° Ø difference exists between the arms of the two potentiometers.

Sideband Switching

In order to switch sidebands (attenuate the lower and transmit the upper) all that need be done is to add or subtract 180° of phase shift to either audio input of fig. 4.2-B. In the Central Electronics 10A Multiphase exciter, the double balanced modulator is driven by a pair of transformers. The USB-LSB switch simply reverses one of the transformer windings.

Other Transmission Modes

The phasing exciter (either fundamental or heterodyne type) is capable of transmitting several different types of modulation. As an example, it is possible to transmit a single sideband (either upper or lower) with carrier, merely by changing the setting of one of the carrier balance potentiometers. The degree of carrier injection will be determined by the amount of unbalance.

Ordinary double sideband with carrier transmission may be generated with a phasing exciter. To accomplish this, one of the balanced modulators must be disabled, and the carrier balance of the other modulator "upset". The amount of carrier injection must be sufficient to provide proper demodulation of the sidebands at a distant receiver. There is one good point about producing AM this way. If over modulation takes place, negative peak clipping does not occur, therefore, there are no spurious splatter products generated. However, there is second harmonic distortion present, but a fair amount of this may be tolerated before the signal becomes unintelligible. If the carrier is not inserted, the popular double sideband, suppressed carrier (DSBSC) form of transmission will result.

Phasing exciters can also be made to produce phase modulation. Phase modulation is generated by having a double sideband suppressed carrier signal (as

The Lakeshore "Phasemaster" II-A exciter generates the sideband on a convenient frequency (9.0 mc) and then heterodynes it to the desired Amateur band. The circuitry for the double balanced modulator is shown in Fig. 4.2B.

produced in either of the balanced modulators) combined with a carrier that is shifted in phase by 90° from that originally present in the balanced modulator. This can be accomplished in the phasing exciter by taking the sidebands with no carrier from one balanced modulator and adding a carrier with no sidebands from the other balanced modulator. Switches in the modulation circuit can connect the various signals to produce any of the above transmission modes.

4.3 FUNDAMENTAL PHASING EXCITERS

As stated earlier, elimination of the unwanted sideband can be accomplished at the operating frequency. Such a system is used in the Heath SB-10 Sideband Adapter and a simplified circuit of the phasing section is shown in fig. 4.3-A.

Driving voltage from an r-f source is applied to a broadband phase shift network. This configuration is known as a passive type, and is discussed in the section devoted to r-f phase shift networks. The r-f phase shifter is capacitively coupled to the balanced modulator grids. Again, in this configuration of double balanced modulator, there are two identical section. The input to V1-A-B is in-phase, as is the input of V2-A-B. The plates, however, are connected in push-pull and cancellation of the carrier occurs in the output circuit. Four-phase audio signals are applied to the balanced modulators so as to cancel the unwanted sideband, while retaining the wanted sideband. The cathodes of the balanced modulators are returned to ground through balance or null controls. Sideband switching is accomplished by reversing the audio leads to either balanced modulator. R-F chokes in the audio leads are necessary to prevent the audio transformers and associated circuitry from loading down the r-f section.

This circuit may be used to drive a final amplifier directly, but good engineering practice dictates the use of a buffer stage. Since it is not necessary to heterodyne the sideband signal, no frequency conversion stages need be included.

It is virtually impossible to maintain the 90° r-f phase shift at all frequencies and other components must be switched into the input circuit when changing bands. Phase shift errors may even occur near the band edges. The Johnson Pacemaker overcomes this difficulty by using a circuit similar to fig. 4-3-A, but with a crystal controlled signal source. Separate crystals are used for each band and the output is heterodyned to the desired operating frequency.

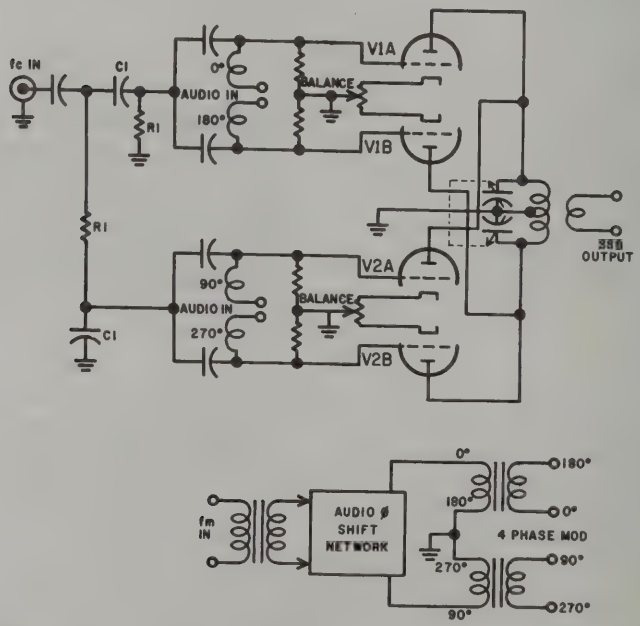


Fig. 4.3A—A double balanced modulator for generating a single sideband. This system is used in the Heath SB-10 sideband adapter.

The Johnson Pacemaker is unique in that the single sideband is generated at a different, and fixed, frequency on each Amateur band. Crystal controlled oscillators drive a circuit similar to that shown in 4.3A, and the output is heterodyned to the desired frequency.

4.4 EFFECT OF PHASE AND AMPLITUDE ERRORS

The ability of the equipment to maintain the 90° phase difference, and amplitude equality, will determine to what extent the unwanted sideband is suppressed. It has been shown that if the two audio voltages differ in phase by an angle A and assuming that all other circuit conditions are perfect, that the suppression of the unwanted sideband with relation to the wanted sideband will be:

$$\text{Suppression in db} = 20 \times \text{Log} \times \text{Cot } A/2$$

If actual values for angle A are assumed and substituted in the above formula, we find that a phase error of one degree will result in a sideband suppression of 40 db, an error of two degrees will produce 35 db, 3.5° will result in 30 db of suppression and so on. This will give the reader an idea of how much an error in the phase shift network will affect the output signal. Again, this was assuming that the amplitude relation was perfect and also that the r-f phase shift network is perfect.

Errors in amplitude equality of the two audio signal voltages will also affect the amount of sideband suppression. The amount of effect is expressed by the formula:

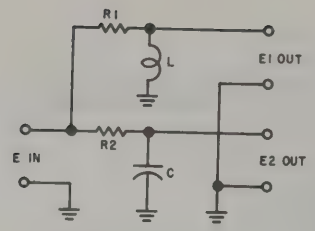
$$\text{Suppression in db.} = 20 \text{ Log } \frac{200 + E}{E}$$

where E is the difference in the two voltages expressed in percent. Thus a difference between the two voltages of 1% would result in a sideband suppression of 45 db, an error of 2% would yield 40 db, 4% results in 34 db suppression, and so on.

4.5 RF PHASE SHIFT NETWORKS

There are various methods of obtaining r-f phase shifts, but the easiest method is that of using two lightly coupled tuned circuits. One of the circuits is detuned on the high frequency side of resonance, to the three db point. This is the frequency where the voltage across the tuned circuit is three db (or 70.7%) less than the voltage when tuned to resonance. The other tuned circuit is set at the three db point on the low frequency side of the carrier. Under these conditions, the voltages existing across the two coupled link windings (see Fig. 4-2-B) are 90° apart in phase relationship.

The phase and amplitude accuracy of any r-f network will affect the sideband suppression directly. The same formulas that describe sideband suppression versus errors of phase and amplitude can be used for r-f phase shift networks as well as audio phase shift networks.



f (mc)	R1-R2 (ohms)	L (uh)	C (mmfd)
3.9	50	2.04	817
7.25	50	1.10	439
14.25	50	.56	223
21.2	50	.37	150
29.0	50	.274	110

Fig. 4.5A—An R-L-C radio frequency phase shift network and component values for the popular ham bands.

The R-L-C Network

Fig. 4.5-A shows the schematic of a simple r-f phase shift network which will perform satisfactorily if the components are chosen properly. The phase relationship is always 90° regardless of the frequency if the component sizes satisfy the following relationships:

$$C = \frac{1}{2\eta F_n R} \quad \text{and} \quad L = \frac{R}{2\pi F_o} = \frac{R}{2\eta F_n}$$

The resistances of R1 and R2 are equal and will determine the impedance of the network. F_n is the center frequency of each Amateur band.

However, the amplitude characteristic of the network will not be constant. It will be seen that if the frequency is moved *n* percent from the design center frequency, the amplitude difference between the two output voltages will be *n* percent. Knowing the possible values of *n*, we can estimate the worst sideband suppression that is due to the r-f phase shift network. The 75 meter band is the widest Amateur band percentage-wise, that is, the 200 kc width divided by the 3900 kc mid-band frequency. Assuming a design center frequency of 3900 kc, the maximum deviation for the vfo to cover the whole phone band would be 100 kc. Thus 100/3900 x 100 = 2.56%. So for 2.56% change in frequency there will be a difference in amplitude between the two output voltages of 2.56% also. From the formulas given in section 4.4, we find that will result in a sideband suppression of 37 db, if all other conditions are perfect.

The input impedance of the network in fig. 4.5-A is resistive and is equal to R (R1 and R2 are equal). The output impedance is very high and should not be loaded down by grid resistors or should not be operated into tube grids that draw current over a portion of the r-f cycle.

If a value is assumed for the two equal resistors to match coax cable from a vfo (50 ohms, as an example), it is then possible to calculate the values of L and C. The table accompanying fig. 4.5-A shows the values calculated for each Amateur band from 80 through 10 meters. Component values for other assumed values of R can be worked out by the interested reader.

The effect of nominal tube input capacitances shunting the inductive branch of the network is very slight and for all practical purposes may be neglected. An attempt should be made to hold any shunting capacity to a minimum. Short leads are a necessity.

The R-C Network

Fig. 4.5-B shows another r-f phase shift network that might be used. For each

f (mc)	R1-R2 (ohms)	C1-C2 (mmfd)
3.9	50	817
7.25	50	439
14.25	50	223
21.2	50	150
29.0	50	110

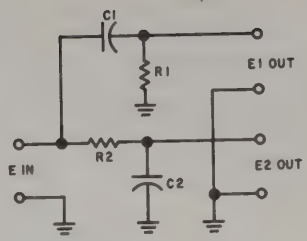


Fig. 4.5B—The R-C radio frequency phase shift network and component values. This circuit is used in the Heath SB-10 adapter.

design center frequency, the capacitive reactance of the two capacitors should be made equal to the value chosen for the two resistances. The input impedance of this network is not a pure resistance and varies considerably with frequency. This might possibly give trouble by causing a varying load on the vfo as the frequency is changed. This circuit is currently being used in the Heath SB-10, to produce the required phase shift. In operation, it appears to have excellent broad band characteristics which results in extremely good sideband suppression even near the band edges.

4.6 AUDIO PHASE SHIFT NETWORKS

This formidable sounding gadget is merely a group of resistors and capacitors chosen carefully and arranged in a certain way, so that the required 90° of audio phase shift will result. This 90° phase difference must be maintained at all the speech frequencies in which we are interested, normally from 300- 3,000 cycles. The other requirement of this network is that the two output voltages must have exactly the same amplitude with respect to each other for all frequencies in the speech range. These are quite rigid requirements, and it was only in recent years that designers have been able to come up with practical networks that could be easily constructed. One percent components are usually incorporated in the network, for as we have seen, any deviation from the network requirements will result in degrading of the sideband suppression.

The Dome Vacuum Tube Circuit

The big "hooker" in the audio phase shift network is maintaining the correct phase angle over the range of speech frequencies. R. B. Dome devised the first widely used circuits. He showed that it is possible to build circuits using RL and C combinations, but more important showed how the results could be obtained by using only RC circuits. This is the easiest approach because of the difficulty in making an inductance that has a low distributed capacity and low resistance. It is easier to build a pure resistance and a pure capacitance. The Dome vacuum tube circuit was originally used by Norgaard, W2KUJ, in his exciter and YRS-1 single sideband receiving adapter. This system successfully maintained the amplitude and phase relationships necessary over the voice frequency range from 50- 5,500 cycles. This more than provides the minimum speech frequency spectrum necessary to carry on satisfactory communications. The Dome system is not found in modern Amateur equipment because of the additional complication of the vacuum tubes. More information on this circuit will be found by locating the references listed at the end of this chapter.

The Passive Network

The most popular form of audio phase shift network employs only R and C in a passive system — no vacuum tubes. There are two popular passive networks and both are manufactured commercially. Both units are designed to work between 300- 3,000 cycles, which is considered adequate for good intelligible speech. The same configuration is employed in each network, however the component values differ because of the design impedance.

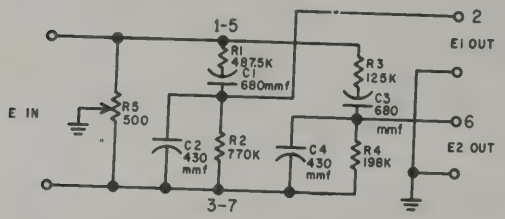
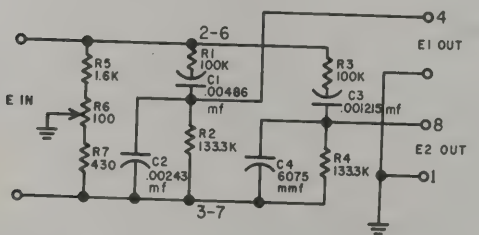


Fig. 4.6A—Complete passive audio phase shift network for 300-3000 cycles. Circled numbers indicate pin connections on the B&W 2Q4 network.

C1, C3—680 mmfd., ± 2% silver mica
C2, C4—430 mmfd., ± 2% silver mica
R1—487,500 ohms, ± 1% deposited carbon, ½w. (IRC)
R2—770,000 ohms, ± 1% deposited carbon, ½w. (IRC)
R3—125,000 ohms, ± 1% deposited carbon, ½w. (IRC)
R4—198,000 ohms, ± 1% deposited carbon, ½w. (IRC)
R5—500 ohm potentiometer, linear taper (Note: this control is not included with the commercial network)

The network shown in fig. 4.6-A is designed for a 500 ohm input impedance and is currently being marketed by the Barker & Williamson Co. (Type 2Q4—Model 350). The network shown in fig. 4.6-A uses 2% tolerance capacitors and 1% tolerance resistors. The maximum phase error between the two outputs is no greater than 1.5°. This would account for a maximum possible sideband suppression of 37.5 db at the worst possible frequency. The figure would probably be degraded slightly by the possibility of errors in the r-f phase shift network or inequalities in the amplitudes of the audio or r-f signals fed to the balanced modulators.



C1—4860 mmfd., ± 1%, silver mica
C2—2430 mmfd., ± 1%, silver mica
C3—1215 mmfd., ± 1%, silver mica
C4—607.5 mmfd., ± 1%, silver mica
R1, R3—100K, ± 1%, ½w.
R2, R4—133.3K ± 1%, ½w.
R5—1.6K ± 5%, ½w.
R6—100 ohm potentiometer, linear taper.
R7—430 ohms, ± 5%, ½ w.

Fig. 4.6B—Passive audio shift network used in the Lakeshore and Central Electronics "Multiphase" exciters. Resistors R5, R6, and R7 are external to the commercially available networks. Circled number indicate pin connections on the Central Electronics PS-1.

The circuit configuration of fig. 4.6-B is the same as the audio phase shift network used by Norgaard in his *SSB Jr.*, in the Central Electronics line of Multiphase exciters, and by Lakeshore Industries. The performance of either circuit is identical, only the component values and the driving impedance is different.

Although component values are given for Fig. 4.6-A and 4.6-B, it will be found that the commercial networks may be purchased for less than the cost of construction. Audio phase shift networks are available from Barker & Williamson, Central Electronics, and Millen Co.

Network Input and Output Impedances

The networks shown in fig. 4.6-A and 4.6-B *must* be properly matched in order to perform properly. The input impedance of fig. 4.6-A is 500 ohms and is determined by the 500 ohm balance potentiometer which is in parallel with a fairly high impedance source. The network of fig. 4.6-B is designed to operate from a 2500 ohm source impedance. In either network configuration, they must "look into" an infinitely high impedance load, in practice a pair of class A amplifier grids. If these conditions are not met it will not be possible to obtain a constant 90° phase shift, and the sideband suppression will suffer accordingly.

The 500 ohm potentiometer, shown in fig. 4.6-A will not be at the midpoint position when properly adjusted. This is because the insertion loss of the two halves of the filter are not identical. Although the input is far from being balanced, the output signals will be exactly the same, when the potentiometer is correctly adjusted. This control should be set up at a frequency of 1,000-1,500 cycles so that the output voltages are at 90° relation. The audio oscillator signal is first applied to *both* the vertical and horizontal oscilloscope inputs. The vertical and horizontal gain controls are adjusted so that the trace is at an exact 45° angle on the oscilloscope screen. Then the two scope leads are transferred to the output of the phase shift network and the audio oscillator is connected to speech amplifier input. The oscillator output must be less than .05 volts and the audio gain on the exciter should be set well below the overload point. When the balance potentiometer is correctly set, the trace on the oscilloscope will be a perfect circle. This control can be set "on the nose" for best sideband suppression as described in Section 4.7, Adjusting the Exciter. The components used in the phase shift network are fixed, and no adjustment is necessary.

Restricting the Speech Spectrum

Since the two passive networks are phase accurate over a 10- 1 range (300-3,000 cycles) measures must be taken to restrict the speech frequencies to this range. Any speech frequencies outside the 300- 3,000 cycle limits that pass through the phase shift network will come to the dual output terminals without the necessary 90° relationship. This means that unnecessary spectrum space is being used both on the wanted sideband side of the carrier frequency and on the unwanted side. Thus on the suppressed side of the carrier there will probably be satisfactory suppression from 300- 3,000 cycles and unsatisfactory suppression beyond this range. The human voice contains high frequency components well beyond 3,000 cycle and it is this component that causes the annoying "buckshot" sound heard all too often on the ham bands. Harmonic distortion in the speech amplifier stages can also produce audio signals higher than 3,000 cycles.

It is necessary to insert a low pass audio filter in the speech system preceding the phase shift network. This filter must effectively attenuate the speech frequencies above 3,000 cycles. This can usually be accomplished by a single section balanced *pi* filter of the type shown in fig. 4.6-C. This filter must have a low characteristic impedance so that it can work into the low input impedance of the network shown in fig. 4.6-A.

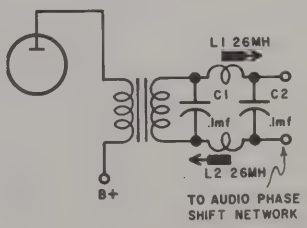


Fig. 4.6C — Speech frequency low-pass filter with a cut-off frequency of approximately 3000 cycles.

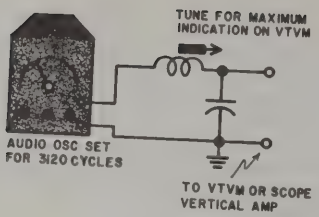


Fig. 4.6D—Test set-up to be used in adjusting a TV width coil for a 26 mh inductance value.

The 26 mh coils of fig. 4.6-C (L1 and L2) can be toroid wound inductances or can be a pair of slug tuned TV width coils. These are available in all television parts stores. The Stancor WC-5 coil has an inductance range of 4 to 39 mh and should be adjusted to the specified 26 mh value by using an impedance bridge or "Q" meter. If such an instrument is not available the slug setting can be made by resonating the coil with a 0.1 mfd. capacitor at 3120 cycles. The layout shown in fig. 4.6-D can be used for adjusting the slug. The oscilloscope or vtvm indication will be maximum for the series resonant condition. The two coils should be mounted so that there is minimum mutual coupling between the two windings. If toroidal coils are used, they may be mounted on the same support because a toroid has no external field. The characteristic impedance of this filter is 500 ohms. A commercial band-pass filter is available for this application from Standard Transformer Corporation (Stancor) and carries the part number C-2340.

It is also necessary to attenuate the frequencies below 300 cycles, but this may be accomplished with ease by using low value coupling capacitors (.001- .005 mfd.) and small cathode bypass capacitors, in the audio stages.

4.7 ADJUSTING THE EXCITER

The fact that amplitude and phase errors in either the r-f or audio networks will degrade the sideband performance, can work to our advantage. It is possible to connect test equipment in such a manner so as to show the quantity of unwanted sideband energy. The various phase and amplitude adjustments can then be set for best sideband suppression. Before proceeding, though, let's review the adjustments that will affect the sideband suppression.

The r-f phase shift network components may, or may not, be adjustable. As we have seen, the phase shift must be *exactly* 90° between the two channels. It would be extremely difficult to use an oscilloscope directly to set up this section. Not only would the oscilloscope test leads disturb the tuning of the coils, but few Amateurs possess oscilloscopes that respond to frequencies of 9 mc or higher. The energy in the r-f phasing circuit is not sufficient to drive the deflection plates directly. This circuit must be set, then, for best sideband suppression.

The audio gain control will determine the amount of r-f energy delivered by the exciter to the load. When testing or operating, be sure this control is set well below the overload point. If the audio stages are overdriven, the harmonic distortion created will completely confuse and invalidate the results.

The a-f phasing potentiometer is located at the input of the audio phase shift network and determines the ratio of signal fed to the two network channels. This ratio is not 1:1, but more like 2:7 because of the unequal insertion loss of the two network sections.

The a-f balance control is usually located in the cathode circuit of the tubes driving the balanced modulators. Proper adjustment of this control insures that the two signals leaving this stage are *exactly* equal in amplitude.

The carrier balance potentiometers are located in the balanced modulator circuit. When these potentiometers are correctly set, there will be no carrier energy in the exciter output. When tuning up a phasing exciter the carrier must be balanced out so that it will not be confused with the unwanted sideband energy.

Equipment

To do an alignment job on a phasing exciter you will need an audio oscillator and an oscilloscope. The oscillator waveform must be quite pure and contain less than 1% distortion. The Heath AG-9 is an excellent piece of test equipment for this application. The calibration accuracy should be at least 10%, that is, 100 cycles in 1000 cycles. It is necessary to know the frequency within 100 cycles or so. The B&W network (Model 2Q4) requires a 1000 cycle test signal and the Central Electronics unit (Model PS-1) is aligned with a 1225 cycle tone. If the audio signal generator does not have a vernier output attenuator, a voltage divider consisting of a 100K and a 10K resistor should be used between the oscillator and the exciter input.

The oscilloscope is used to observe the quantity of unwanted sideband energy. The exciter output is connected to a resistive dummy load. Do not be tempted to use a lamp for the load, since heating of the element will change resistance, thereby varying the load on the exciter. R-F energy must be applied directly to the deflection plates in most oscilloscopes. To obtain sufficient voltage to scan the beam, it is necessary to employ a resonant set-up coil as shown in fig. 4.7-A. A Transitron "Q" probe is an excellent piece of test equipment for this application.

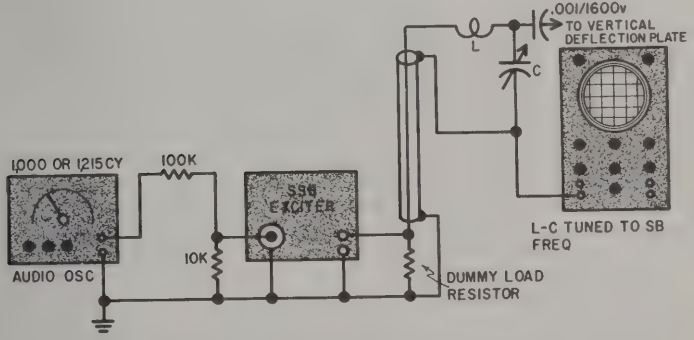
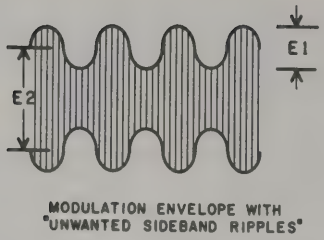


Fig. 4.7A — Typical test equipment set-up for phasing exciter alignment.

The theory of adjusting the exciter for best suppression is easy to comprehend, if you recall the basic principle of SSB. Apply a *single* tone to the exciter, say 1000 cycles. If *everything* is perfect what will result? The output energy will consist of a *single* frequency which is displaced from the suppressed carrier. For all practical purposes, the energy is the same as would be obtained from a cw transmitter. On an oscilloscope, the cw energy will appear as a square or rectangle (depending on the gain settings) of light. More important, the top and bottom of the pattern will be *flat*, indicating the complete absence of modulation. If the cw transmitter power supply is not properly filtered, hum modulation will result, and 120 cycle ripple will appear on the top and bottom of the oscilloscope pattern.

The same indications apply when adjusting a single sideband exciter. If the only energy applied to the dummy load was the single sideband, no ripple would appear on the oscilloscope pattern. Since this condition seldom exists, the oscilloscope



$E_1 \cdot E_2$ Ratio	Suppression
1:10	20 db
1:15	24 db
1:20	26 db
1:30	30 db
1:50	34 db
1:100	40 db

Fig. 4.7B—Sideband suppression can be gauged from this chart. Note that a 1:100 ratio is pretty rough to observe and measure on an oscilloscope.

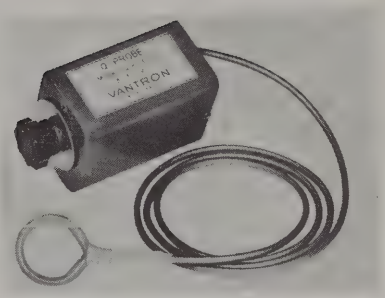
screen in fig. 4.7A will display a slight amount of ripple. The amount of wiggles on the trace will depend on the settings of the exciter adjustments. The ratio of ripple amplitude to average indicates the amount of sideband suppression. The chart in fig. 4.7-B shows the degree of suppression you can expect for various patterns.

Both unwanted sideband and carrier energy will appear in the form of ripples. It is easy to distinguish between the two, however, for the unwanted sideband ripples will be half as wide as the carrier ripples.

Once the set-up shown in fig. 4.7-A is completed, the r-f phase shift adjustments audio phase control, audio balance control, and carrier balance controls should be set for minimum ripples on the top and bottom of the scope pattern. For fundamental phasing type exciters, the vfo should be placed at the center of the band. With some exciters it will be necessary to adjust the individual r-f phase shift networks on each band. If the pattern is devoid of wiggles, you're kidding yourself brother! Either the exciter is tuned to the vfo (a cw source) or one of the exciter stages is overloaded. It will be noted that many of the controls interact and the adjustment of each control should be repeated several times. It may also be noted that when the suppression of a particular sideband is best, the suppression of the other sideband may not be optimum. This is a characteristic of phasing exciters and must be "lived with". If operation is to be confined to the three hf bands, the upper sideband can be favored. If all band operation is contemplated, a compromise setting is in order. The technician should switch sidebands after each adjustment is made. A compromise setting will be one in which the suppression is as good as possible and approximately equal on each sideband. Once you are satisfied with your alignment job, check the suppression at each end of the audio range (300 and 3000 cycles). The wiggles will increase, but it will not be too severe if the equipment is properly adjusted. If the exciter being aligned is a fundamental phasing type, it might be informative to check the suppression at the band edges. In general this system may be used to evaluate carrier suppression in filter type exciters, but the suppression adjustments will not be used, or will not be the same. For more detailed alignment procedure, such as the tuning of heterodyne stages and the placement of harmonic traps, refer to your instruction manual.

A properly adjusted phasing exciter will exhibit approximately 35 db of sideband suppression. The "red blooded" experimenter may desire to improve this figure. The improvement can usually be had by varying the components associated with the audio phase shift network. You must be sure that the equipment is operating properly, and that the exciter is producing the best possible suppression, before attempting any modifications.

Another interesting method for checking and adjusting the exciter is with the station communications receiver. The receiver must be equipped with a crystal filter or stable "Q" Multiplier so that it will be possible to tune in the various components of the sideband signal. The carrier balance controls will null the carrier



The Vantron "Q" Probe is an ideal instrument for oscilloscope measurements of sideband exciter performance. The device uses a remote pickup loop that is coupled to the loaded output coil. The link drives a multiband tuner that can be resonated between 3-30 mc which provides the necessary voltage step-up for the oscilloscope plates.

frequency and the other exciter adjustments can be set for minimum unwanted sideband energy. Needless to say, the receiver must be very stable and it will be necessary to tune back and forth between the two sidebands to insure that you are not suppressing the wanted sideband also! Alignment with this system can be extremely tedious and quantitative measurements are subject to the whims of the "S" meter. Although the oscilloscope system requires more test equipment, and is only informative to 40 db (or so), it is very satisfactory for alignment jobs on Amateur equipment.

4.8 THE HETERODYNE PHASING EXCITER

The author is grateful to W9DYV of Central Electronics for permission to publish a portion of his 10A "Multiphase" exciter schematic. This very popular unit is available commercially, either in kit form or completely wired.

The circuit is basically that of the "SSB Junior" devised by Norgaard, but with improvements that make multi-band operation possible. The original SSB, Jr. exciter was a fundamental frequency operating gadget. It operated at 4.0 mc and if the operator wished to QSY more than a few kilocycles, he found it necessary to realign the r-f phase-shift network in order to maintain good sideband suppression. Wes, W9DYV, modified the idea by generating the SSB signal at a fixed frequency of 9.0 mc and heterodyning into the desired Amateur band with a separate mixer stage, just as was done in the filter type exciter. VFO operation is, of course, possible when using this scheme.

Remembering the foregoing discussion about the phasing method of generating SSB signals, we now refer to fig. 4.8-A. V1a and V1b are the usual speech preamplifiers. V2a is the a-f driver stage which feeds the audio phase-shift network through the transformer, T1. Construction by the average ham of the a-f phase-shift network is possible, but a completely aligned unit is available for about the same price for which one could buy the necessary precision stable components.

V3a and V3b are the dual channel amplifiers and the audio balance is obtained by adjustment of the cathode resistor, R18. The transformers, T2 and T3, are plate to low impedance line transformers used to drive the diode balanced modulators. W9DYV indicates that suitable units are not easily obtained on the market so he has special transformers built for his production needs. Switch, S1, is the function switch which permits selection of sidebands or of AM or PM transmission. Studying the switch positions will show that changing from one sideband to the other merely interchanges the connections on the output of T2, while switching to the AM and PM positions disconnects the secondary of T3 from its associated balanced modulator.

The audio in the two separate channels is applied, in series with the r-f from the link windings on L1 and L2, to the balanced modulators. Y1 and Y2 are the

diodes of the balanced modulator involved with L1 and T2. Likewise Y3 and Y4 are associated with L2 and T3.

The R-F Circuitry

V2b is the 9.0 mc crystal oscillator, and utilizes L1 as its plate tank coil. The resonant frequency of the L1/C11 combination must be higher in frequency than 9.0 mc, in order for the crystal to oscillate; therefore, to obtain our 90° r-f phase-shift, L2 and C13 will have to be tuned lower in frequency than 9.0 mc. You will notice that there is no physical connection from L2 to the oscillator tank. L1 and L2 may be mounted physically within a couple of inches of each other, and the circuit capacity furnishes sufficient coupling to do the job. Capacitors C9 and C10 prevent r-f from getting into the audio transformers; yet they must not appear as a low reactance at audio frequencies.

Carrier balance is accomplished by adjusting R23 and R24. Both must be carefully balanced in order to cancel the carrier completely.

The balanced modulator output transformer L3 is the point where the outputs of the two previously separate channels are combined. The additional double tuned transformer, consisting of L4 and L5, is necessary to further attenuate the second harmonic of the 9.0 mc oscillator, generated in the germanium diodes of the balanced modulator stage.

The mixer stage, V4, is conventional with the exception of the trap circuits, L6 and L7, and their associated tuning capacitors. They are necessary, when operating in the 14 mc Amateur band, to attenuate the third harmonic of the 5.0 mc mixing voltage. The same vfo may be used for both 4.0 and 14 mc operation, for the difference between the 9.0 mc SSB signal and the 5.0 mc mixing voltage will put the output at 4.0 mc. If the sum-mixture is selected by the mixer plate tank circuit, L8 and C27, the output will be in the 14.0 mc Amateur band.

Parts List for Fig. 4.8A

C1.C3—.005 mfd., 500v., paper
C2—8 mfd., 450v., electrolytic
C4—250 mmfd., mica
C5—2,430 mmfd., (.002 mfd., mica ± 5% with 170-780 mmfd., trimmer in parallel).
C6—4,680 mmfd., (.0043 mfd., mica ± 5% with 170-780 mmfd., trimmer in parallel).
C7—1.215 mmfd., (.001 mfd., mica ± 5% with 50-380 mmfd., trimmer in parallel).
C8—607.5 mmfd., (500 mfd., mica ± 10% with 9-180 mmfd., trimmer in parallel).
C9, C10, C12, C29, C35—.001 mfd., mica
C11, C13—150 mmfd., mica
C14, C15, C16—.005 mfd., mica
C17, C18—.001 mfd., ± 1% mica
C19, C20—150 mmfd., ± 5% mica
C21, C22, C26, C28, C31, C32, C33, C34—.006 mfd., mica or ceramic
C23—36 mmfd., ± 5% mica
C24—3 mmfd., ceramic or mica
C25—25 mmfd., mica
C27, C36—380 mmfd., air variable
C30—.1 mfd., paper
C37—.25 mfd., paper
PS1—audio phase-shift network (Central Electronics PS1). This unit contains C5, C6, C7, C8, R7, R8, R9, R10.
L1, L2—25 turns, #22 Enamel. Link 4 turns of plastic covered hook-up wire. Closewound on ⅜" form.
L3—8 turns, #16 enamel. Link 1 turn of plastic covered hook up wire. Closewound on ⅜" form
L4, L5—25 turns, #22 enamel. Closewound on ⅜" form.
L6, L7—18 turns, #22 enamel. Closewound on ⅜" form.
L8—75 meters: 46 turns, #20, wound 16 turns per inch, 1"diameter (B&W 3015). 20 meters: 12 turns #14 wire, 1⅝" long, 1⅛" diameter.
L9—Same as L8, but tap placed for proper loading, approximately 15% up from "cold" end
NOTE: L1 and L9 may be purchased from Central Electronics Company
R1—1 megohm, ½ w.
R2, R5—2,200 ohms, 1w.
R3—100,000 ohms, 2w.
R4—220,000 ohms, ½w.
R6—220,000 ohms, 2w.
R7, R10—133,300 ohms, 1%, ½w.
R11—10,000 ohms, 1w.
R12—1 megohm potentiometer
R8, R9—100,000 ohms, 1%, ½ w.
R13, R26—47,000 ohms, ½ w.
R14, R19, R21—1,000 ohms, 1w.
R15—400 ohms, 1%, ½w.
R16—1400 ohms, 1%, ½w.
R17, R20, R31—560 ohms, 1w.
R18—1,000 ohm potentiometer
R22—100,000 ohms, ½ w
R23, R24—1,000 ohms carbon potentiometer
R25, R35—1,000 ohms, ½w.
R27—180 ohms, ½w.
R28—4,700 ohms, 2w.
R29, R30—15,000 ohms, 5w.
R32, R38—swamping resistor
R33—56 ohms, 1w.
R34—22,000 ohms, ½w.
R36—560 ohms, ½w.
R37—100 ohms, 1w.
R39—10,000 ohms, ½w.
R40—100,000 ohms, 1w.
S1—3 pole, 4 position wafer switch
S2—d.p.d.t. wafer switch
S3—relay contacts on voice control relay (see text)
T1—interstage transformer (special) Central Electronics Type 27AM-24
T2, T3—singl plate to voice coil (special) Central Electronics Type 27AO-79
NOTE: T1, T2, T3 may be purchased from Central Electronics **Y1, Y2, Y3, Y4**—1N34 Germanium diodes

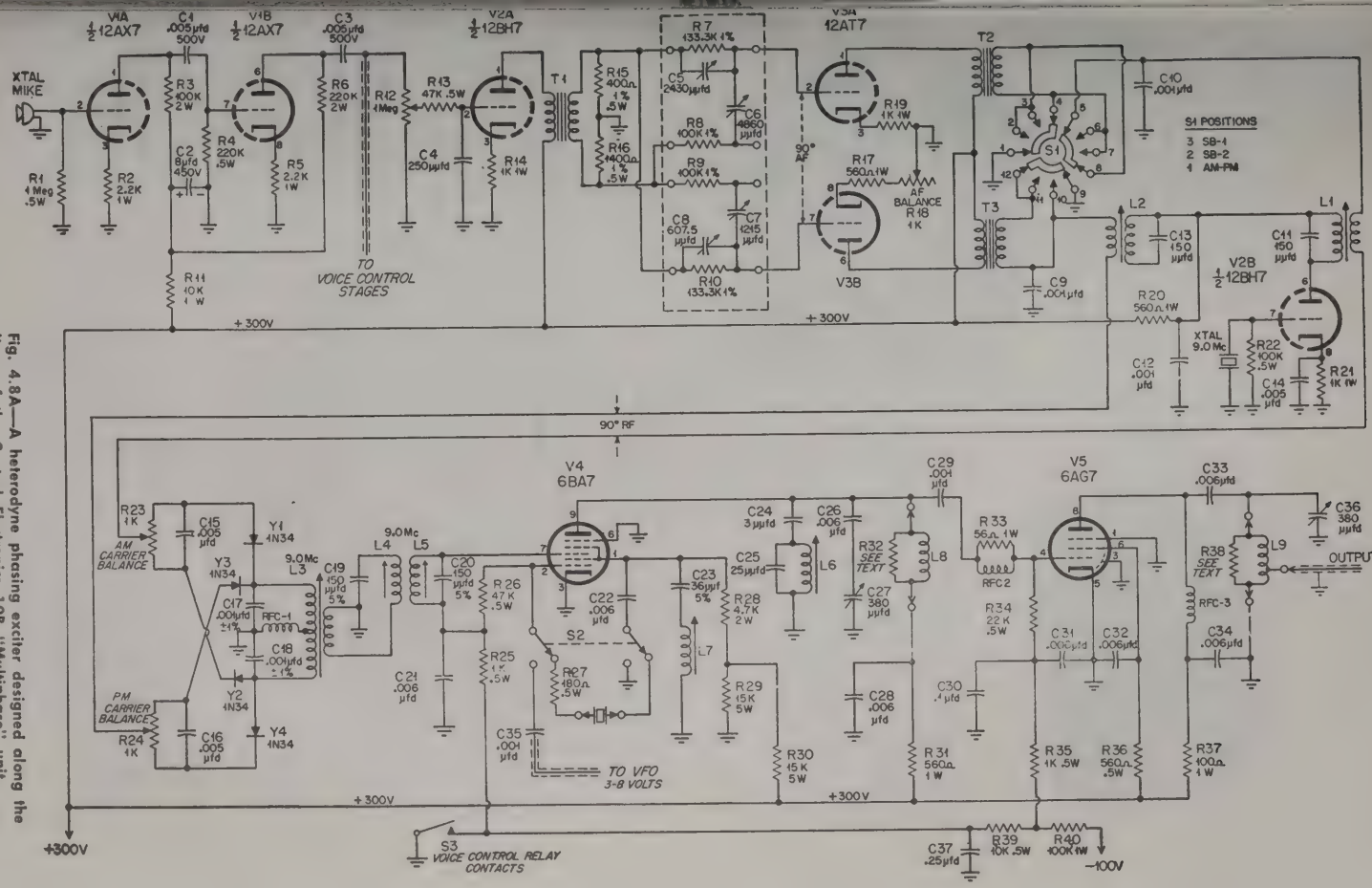
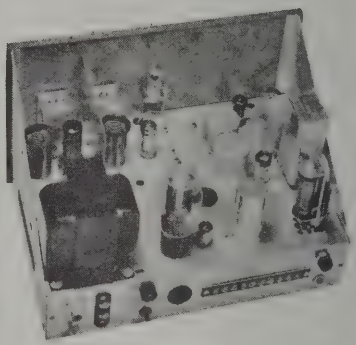


Fig. 4.8A—A heterodyne phasing exciter designed along the lines of the Central Electronics 10B "Multiphase" unit.

The Central Electronics 10B "Multiphase" exciter.



Chassis rear view of the Model 10B "Multiphase" exciter. This unit includes a power supply not shown in the general schematic for this unit. The output stage, V5, with plug-in coils is in the upper left in the shielded section. V4 is in the center of the chassis. The audio phase shift network, PS-1, is mounted in the rectangular can beside the relay at the lower right.

The output amplifier, V5, is the inevitable 6AG7. The output is shunt-fed, and obtained by means of a tap on the plate coil.

The swamping resistors, R32 and R38, are usually necessary in order to stabilize the 6AG7 stage. Without these resistors, self-oscillation often results. The exact values of these resistors are not shown; the highest value of resistance commensurate with stable operation should be used. Always insure that all stages are rock solid before the unit is "buttoned up".

You will note, when in stand-by position (S3 open), that there is a minus 100 volts of bias applied to the control grid of the 6AG7 and the oscillator grid of the 6BA7 mixer. This thoroughly squelches the output of the exciter. When S3 is closed, the mixer returns to its normal operating condition and an operating bias of about minus 10 volts is applied to the control grid of the 6AG7.

Forty-meter operation is also possible with this unit. However, it is not recommended that a mixing voltage at 1.8 mc be used, because of the various harmonics of this frequency that will fall in or near the 7.2 mc phone band. These harmonics are not necesarily present in the vfo output, but are generated in the electron stream of the mixer stage. In view of this, the use of a 16.2 mc vfo is recommended. This might take a little doing, but is not an impossible task. In any case, a frequency stable vfo voltage of about 3 to 8 volts is needed.

Alignment

Adjust the 9.0 mc oscillator tank, L1 for oscillation of the crystal. With an appropriate mixing frequency fed into the vfo jack, and the receiver tuned to the desired mixture-output frequency, adjust L3, L4, L5, L8 and L9 for maximum output with one of the carrier balance pots slightly off balance.

Connect an oscilloscope to the output of the 6AG7 amplifier and use a recurrent sweep rate of about 30 per second. Now carefully balance R23 and R24 for as

perfect carrier balance as possible. Feed a steady tone of about 1000 cps from an audio oscillator into the microphone jack. Make sure that the audio wave-form is good and that nothing is being overdriven. You will see on the oscilloscope pattern that resembles a modulated AM envelope. Next, adjust the audio balance control, R18, and the r-f phase-shift network, L2, for minimum ripples (or modulation) on the oscilloscope pattern. The pattern for a properly aligned SSB exciter with single-tone input is a pure cw envelope with no modulation. The presence of ripple indicates one of three things: (1) Presence of undesired sideband signal; (2) carrier unbalance; or, (3) bad wave-form in the input audio tone or distortion produced by overdriving the audio stages.

4.9 A SINGLE SIDEBAND ADAPTER

As explained earlier, an adapter differs from an exciter in that it contains no carrier source. An r-f signal is fed to the adapter from an external source (such as an existing AM transmitter) and the unit converts it to a single sideband, suppressed carrier signal. The output from the adapter may be used "barefoot" or it can be used to drive the final of the existing AM transmitter. This, of course, assumes that the final has been modified for linear operation.

The Heath SB-10 is such a device. The schematic diagram is shown in fig. 4.9-A and includes the balanced modulator, class A driver and output stages, and VOX/anti-trip circuitry. The SB-10 will deliver 10 watts PEP output with only 3 watts of drive at the fundamental frequency. A pi-network output circuit effectively matches the interconnecting cable. The adapter uses 9 tubes, operates on the 80 through 10 meter bands, and of course generates the sideband with the phasing system. The unwanted sideband suppression is in excess of 30 db, and with proper adjustment, the carrier suppression is in excess of 40 db. The SB-10 is contained in a cabinet 10″ high, 6¾″ wide, and 13″ deep and is complete except for the external power supply. The power requirements are 350 volts @ 85 ma (average) and 6.3 volts @ 3.5 amperes.

Balanced Modulator Circuitry

The input circuit to the balanced modulator pair consists of an untuned broadband r-f phase shift network of the passive type (described in 4.5-B). An entirely different network (to provide the proper phase shift at each fundamental input frequency) is switched in on each band. This permits operation on all frequencies in the particular band desired without the necessity of adjustment, but still maintains precision in the r-f phase shift network. The network is capacitively coupled to the balanced modulator grids. The cathodes of the balanced modulators are returned to ground through 1000 ohm tube balancing or carrier null controls which are adjusted from the front panel. The plates of the tubes are tuned to the operating frequency with a multiband coil and capacitor circuit.

The modulation voltage developed in the a-f modulator is coupled through two 1:1 ratio transformers to the balanced modulator grids. A sideband selector switch is connected in the output leads of the transformers which applies the audio voltage in the proper phase relationships to obtain upper, lower or double sideband (AM) operation. The r-f voltage is isolated from the audio network by use of four 500 mh chokes to couple the audio to the grids of the balanced modulators.

Driver and R-F Amplifier Circuitry

A 6CL6 is employed as a Class A driver stage. The grid of the 6CL6 is coupled through a capacitor to the 4 turn link on the balanced modulator output coil. Bandswitch selected, broadband, slug tuned coils are used for tuning the driver output circuit.

The output circuit of the driver stage is coupled to the grid of the 6BQ5 tube. This stage is also operating in Class A service and is easily driven to maximum output while maintaining excellent linearity. The plate circuit is shunt fed and couples to an all-band pi network output circuit. The plate circuit also furnishes a sample of r-f voltage to operate the meter for the purpose of tuning the adapter.

Audio and Phasing Circuitry

The 12AX7 dual triode tube provides two stages of resistance coupled speech amplification with sufficient gain for a high impedance microphone input. The design of the speech amplifier circuitry favors an audio frequency range of from 300 to 3000 cycles per second. The output of the speech amplifier is capacitively coupled through a gain control to one section of a 12AT7 acting as an audio driver. The output of the driver is transformer coupled to the low impedance side of a 90° audio phase shift network which is designed to operate over the audio range of 300 to 3000 cycles per second. By restricting the audio frequencies to the stated limits, errors in the 90° phase shift network which will directly impair the suppression of the unwanted sideband, are minimized.

A control, known as a ratio balance control, is connected across the low impedance (500 ohms) side of the interstage transformer which feeds the passive network. The function of the control is balancing the input circuit to the network.

The output of the phase shift network is of a relatively high impedance and is connected to the two grids of a dual triode 12AT7 tube. This tube functions as the modulator.

The cathodes of the 12AT7 modulator are biased with a variable and fixed resistor configuration. The purpose of the control is for audio balance. This tube must deliver equal output to the primary of each of the two modulation transformers.

Voice Control and Anti-Trip Circuitry

The output of the speech amplifier is also capacitively coupled to one section of a dual triode 12AT7 tube acting as the VOX voltage amplifier. The other half of the 12AT7 tube acts as the anti-trip amplifier and receives and amplifies the receiver output, which is transformer coupled to the grid.

The output of the VOX voltage amplifier is capacitively coupled to the plate of one section of a 6AL5 dual diode and the outlet of the anti-trip voltage amplifier is capacitively coupled to the cathode of the second 6AL5 diode.

The audio output voltage which is developed in the anti-trip voltage amplifier is then rectified and appears as a negative potential at the plate of the other 6AL5 diode. The diode will not conduct when so biased. Therefore, the relay control tube (½ of a 12AT7) remains cut-off by the fixed bias applied to the cathode by the voltage divided network.

From this anti-trip action, it can be seen that, for the relay to be tripped, it will be necessary to upset the negative voltage at the plate of the diode. This allows it to conduct and place sufficient positive voltage on the grid of the relay control tube to overcome the fixed cut-off cathode bias. This positive "upsetting" voltage is developed in the speech and VOX amplifier stages and originates at the microphone. Thus, with proper setting of the sensitivity controls, speaking into the microphone trips the relay by overcoming the receiver anti-trip bias. A capacitor resistor network charges and discharges across the grid of the control tube to introduce the necessary time delay after speech is discontinued.

A cathode bias voltage divider network on the relay control tube is switched in or out for a selection of voice control or manual operation. In standby position, B plus is removed from all circuits.

Although the SB-10 is primarily designed for converting existing AM transmitters, it may not have occurred to the reader that it is the foundation for an excellent

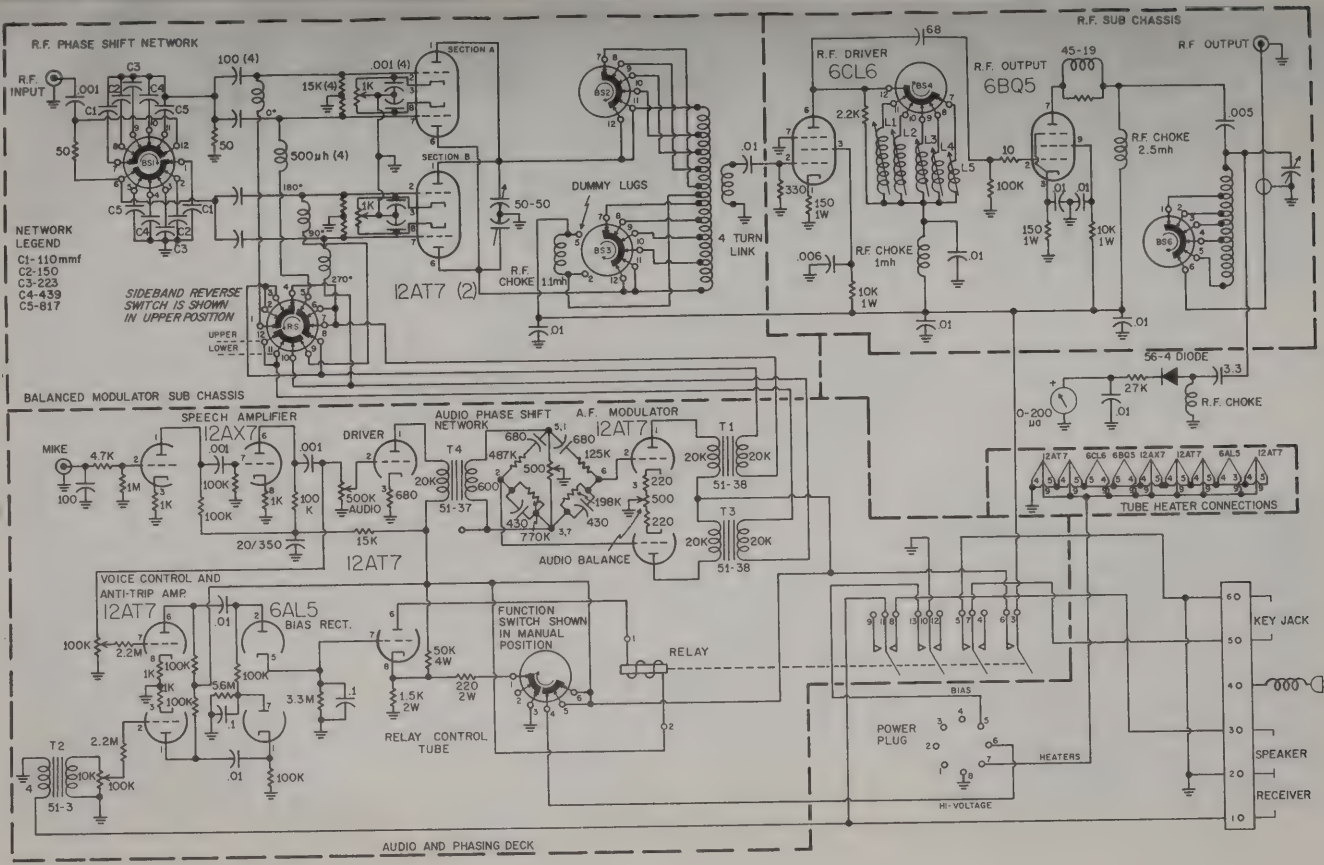


Fig. 4.9A—Schematic for the Heath SB-10 sideband adapter.

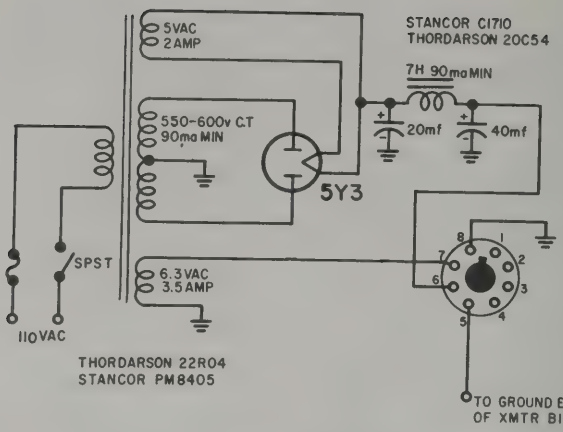


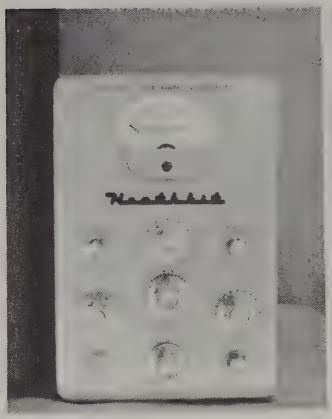
Fig. 4.9B—A power supply suitable for use with the Heath SB-10 sideband adapter.

exciter. Simply by adding a vfo and a linear amplifier, the SB-10 can be used as a complete all band exciter/transmitter.

Three basic changes are necessary when modifying the popular "100 watt table-top" transmitters (DX-100, Viking, B&W 5100, etc.). First, the r-f path must be broken between the output circuit of the driver (5763 in the DX-100) and the grids of the final amplifier. The r-f drive is brought out to a coaxial connector on the rear of the chassis. Also, the final amplifier grid lead is brought out to a second coaxial connector on the rear apron. Second, the final amplifier must be modified for class AB_1 linear operation. The final amplifier screen voltage should be regulated at 210 volts with two OB2 voltage regulator tubes in series. The bias is then adjusted to give 50 ma of resting plate current. Third, provision should be made for opening the ground end of the fixed bias resistor string during standby periods. This will cut-off the final amplifier without removing plate voltage and eliminate excessive switching of the plate transformer primary.

A typical conversion might consist of a DPDT toggle switch mounted on the front panel which is wired to perform two functions. One section would switch the clamp tube in and the VR tubes out of the circuit for class C operation. In the opposite position the clamp tube is switched out, and the VR tubes into the circuit for class AB_1 operation. The 6146 screen dropping resistor becomes the VR tube load resistor in the linear mode. The second switch section would ground the end of the fixed bias resistor string in the normal manner for CW or AM operation (Class C) but placing the switch in the other position would allow this function to be performed remotely in the SB-10. In addition a potentiometer must be added in the bias resistor string as a means of varying the fixed bias to the proper value for class AB_1 operation. Once set, this control will need no further adjustment since the rest of the bias required for class C operation is developed by grid current.

Power for the adapter can be brought out through the accessory socket. If extended periods on SSB are contemplated, it would be a good idea to remove the speech amplifier and modulator tubes to reduce the filament drain. If there is evidence of transformer overloading it is suggested that a separate power supply be constructed. A schematic for a suitable supply is shown in fig. 4.9-B.



The Heath SB-10 is a "fundamental phasing" type exciter. An accessory transmitter provides a source of r-f on the operating frequency. The single sideband, suppressed carrier signal from the SB-10 can be amplified by the transmitter final if it is modified for linear operation.

4.10 A THIRD METHOD OF GENERATING SSB

In addition to the filter and phasing method, there is a "third system" for generating single sideband signals. This system was described in the SSB issue of the IRE Proceedings, Dec. 1956 issue (Weaver). Although the exciter incorporating this system is somewhat more complicated than the usual ham devices, it does have several advantages that makes it attractive to the experimenter. It does not require sharp cut-off filters or wideband phase shift networks. If such an exciter is improperly adjusted (balance or phasing), out-of-channel energy does not appear. Instead, the energy will appear in-channel with the speech inverted.

Because of the low frequency involved, such a system is ideal for transistors. At least one commercial company has produced a SSB transmitter/receiver employing transistors and using the "third method". The Redifon GR-400 is designed for commercial operation on four pre-set channels between 2-10 mc. The power output is 50/60 watts PEP into a 70 ohm line. Fig. 4.10-A is a block diagram for the sideband generation section of the GR-400. The audio is limited to 300- 3300 cycles and is applied to a pair of balanced modulators. In addition, an audio oscillator,

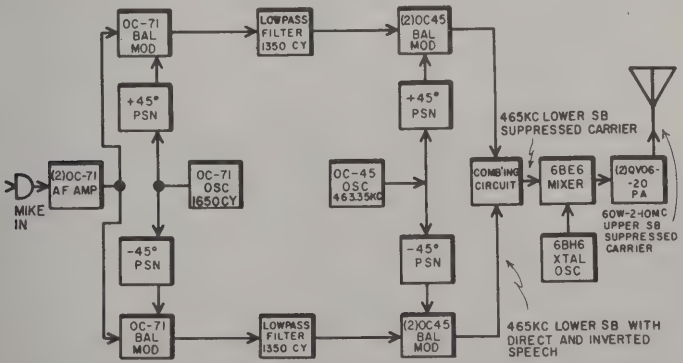
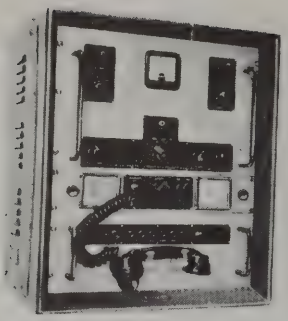


Fig. 4.10A—Block diagram of the Redifon GR-400 transistorized transmitter/receiver. This unit uses the so-called "third method" for generation of the sideband.



The Redifon GR-400 uses transistors in the sideband generation section.

drives each balanced modulator with a 1650 cycle sine wave, through two audio phase shift networks. These phase shift networks are plus and minus 45°. Note that they are not wide band, but "tuned" to the oscillator frequency. The output of the first two balanced modulators consists of two sidebands: one extending from 0-1350 cycles and the other from 1950-4950 cycles and displaced from each other by 90°. A pair of low pass filters are used to remove the high frequency sideband, leaving the 0- 1350 component intact. This sideband drives another pair of balanced modulators and heterodynes with a 463.35 kc carrier oscillator. The carrier is balanced out in the combining circuit which leaves only the lower sideband of a suppressed carrier frequency (463.35 kc). This signal is applied to a vacuum tube mixer, along with a crystal oscillator signal, and is heterodyned to the desired frequency range. Two QVO6-20's are used in the final.

The GR-400 modulator illustrates the "third system" very well. Transistors are used because of the great saving in heater and plate power. Mullard transistors are used but a similar device could be constructed using the American counterparts. For more information on the "third system", the reader is referred to the IRE issue mentioned earlier and to the Sept. 1957 QST article by Howard F. Wright, W1PNB.

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5

Double Sideband

IT HAS been stated repeatedly that the carrier, which we associate with an AM transmitter, adds absolutely nothing to the signal in the way of intelligence. Its only purpose is to act as a reference for demodulation and to make life miserable for the people who have to listen to it. In an AM transmitter, half the power is wasted to apply a signal to the detector that could have been generated in the receiver originally! Pretty silly, isn't it? This fellow traveler is so weak by the time it reaches the antenna that the signal from a miniature tube (or even a transistor) makes an excellent replacement. Rather than aid in the transfer of intelligence from one point to another, the evil carrier hinders the process. We are all well aware of what happens when two carriers "get together". It is much the same as a meeting of two "blow-hards". They don't do much of anything, except toot their own whistles! And the low end of 20 meters usually sounds like a convention of steam calliopes.

5.1 WHY DOUBLE SIDEBAND?

A double sideband, suppressed carrier transmitter is nothing more than a standard AM rig with the carrier subdued to the point where it cannot create interference. There are many advantages to be gained by transmitting DSB, as pointed out by John Costas, W2CRR (Jan. 57, CQ Mag., p. 26). In this system a balanced modulator is also used to create the two sidebands and suppress the carrier. Any of the circuits shown in Chapter II are applicable. However, since the output of the balanced modulator is used "as-is", there is no need to tack on devices to eliminate one sideband. This means that the balanced modulator can be a final amplifier stage, with its output connected directly to the antenna.

Although you may not agree on the relative merits of DSB versus SSB, it does have several advantages over other systems, that cannot be denied. Several of the factors that account for the snowballing popularity of DSB are: (1) elimination of the carrier, (2) more talk power, (3) less average dc power required, (4) speech clipping can be used, (5) a complete lack of critical components, and (6) no carrier phase distortion. Let's examine these points in detail to see why DSB is becoming so popular.

Elimination of the Carrier

This, of course, is DSB's claim to fame. Anyone who has operated the SSB portions of the various ham bands can tell you what a utopia it is without all those nerve shattering whistles to contend with. Since the carrier is suppressed, heterodynes are eliminated. Any "carrier-less" transmitter (even if it uses balanced stomach pumps for carrier suppression) will be welcome on the ham bands.

More Talk Power

All the energy that was needed to generate the useless carrier can now be applied to the sidebands, and this means more talk power. The "juice-hog" modulator is no longer required, and the power consumed in this stage can be utilized by the final amplifier. As we have seen, it is quite practical to apply at least twice the normal plate voltage to a DSB final, without exceeding the plate dissipation ratings.

Less Average DC Power Required

The DSB final amplifier "works" only when audio is applied to the screens. During the peaks of the audio cycle, the final produces peak power. This means that the average power required from the dc supply is far less. In other words, the power supply operates on a short duty cycle. It is permissible to draw more than twice the rated current from the supply, during the modulation peaks, without damaging it. Or, conversely, for a given peak power input the power supply can be physically and electrically smaller than its AM counterpart.

Speech Clipping Can Be Used

Since the output waveform is a replica of the modulating waveform, speech clipping may be employed to increase the average intelligence power (talk power, again). If speech clipping were employed in an SSB transmitter the net effect would not significantly reduce the peak to average power ratio. Since our DSB system still contains both sidebands, this effect does not concern us. The improvement in readability, by using speech clipping, is rather startling. If you doubt this, listen to a Globe DSB-100 with and without clipping. It is so effective that it will make the difference between a Q5 contact or a "sorry OM, but can't hear you anymore" report.

Lack of Critical Components

This is the feature of DSB that appeals to hams all over the world. The

circuitry is simplicity itself, and it is unnecessary to send to the States for a phase shift network, special impedance matching transformers, and so on. Even the modulation transformer can be had by removing the power transformer from an old radio! The famous Command Set transmitter is available all over the world, and it only takes a few hours to convert it to DSB. If your old fashioned AM rig uses a pair of tubes in the final (either push pull or parallel) it can be converted to DSB in an afternoon. That evening you will be "one of the boys"! Single tube final amplifiers can be modified by adding another "jug" and a suitable filament transformer.

No doubt you are already sold, but there are other advantages worth mentioning. Selective fading (or phase distortion) is eliminated because the carrier is no longer around to be confused. Frequency diversity is inherent in the DSB system. The operator has his choice of the more readable of two sidebands. By this time you probably have the transmitter out of the case, so let's not dwell on these advantages, but progress to how all these marvelous things are possible.

5.2 HELP STAMP OUT CARRIERS

A double sideband transmitter is far simpler to build and operate than an SSB transmitter. In fact, a DSB rig is even less complicated than an AM transmitter! Special tricks or gimmicks? Not at all. Just the proper combination of some old and well known techniques. No linear amplifier, no filter, no phasing networks, no frequency translators; an ideal project for the "do it yourself ham". How is all this possible? Well, it is due to the simple fact that we no longer have to transmit a carrier. To see how all these nice things come about, take a squint at fig. 5.2-A.

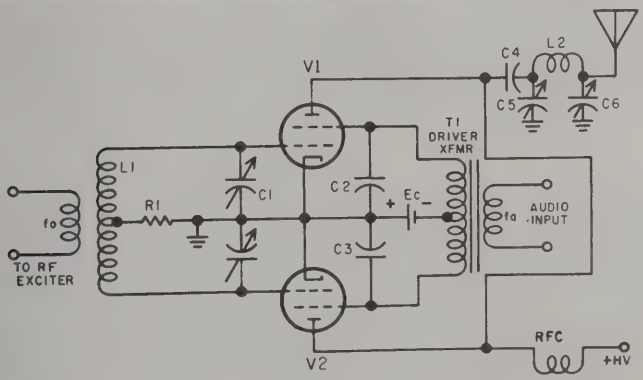


Fig. 5.2A—Schematic of a typical double sideband final amplifier. Note that E_c is often used with high power tubes to avoid exceeding the maximum plate dissipation, and to provide better cross over linearity.

The final tubes V1 and V2 are beam tetrodes and are operated as screen modulated class C amplifiers. The plates are paralleled and are connected to the antenna load by means of a pi network. The control grids are driven push-pull from a normal r-f exciter at the operating frequency. The screen grids are by-passed to ground by C2 and C3 and are connected to the audio modulation transformer T1. A normal driver transformer will handle more audio power than will ever be required for Amateur service. The center-tap of T1 is either grounded or connected

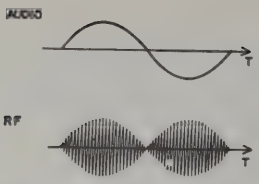


Fig. 5.2B — DSB waveforms, the applied audio (a) and resultant r-f output (b).

to a negative bias supply depending on the tube type and the plate voltage used. Blocking capacitor C4 is used to isolate the plate voltage from the pi coupler.

With no audio, both tubes are nearly cut off by virtue of the fact that the screens are either grounded or biased negatively, thus no output. If we assume a sinusoidal audio tone as the modulating signal as shown in fig. 5.2-B(a), one screen is driven positive during the first half-cycle and the other is driven negative. The tube with the positive screen conducts and r-f is supplied to the load by that tube. During the next half of the audio cycle the other tube supplies the power and the first tube rests. Note that only one tube is working at any one time, except when there is no audio, then both tubes loaf. Fig. 5.2-B shows the audio and r-f waveforms. Only one audio cycle is shown. Note further that the r-f during the first half of the audio cycle is phased 180° with respect to the r-f during the second half of the audio cycle. This is typical of a suppressed-carrier AM signal. Suppose we add a carrier wave to the r-f wave. If the carrier wave has the same phase as the r-f in the first audio half-cycle and an amplitude equal to the maximum amplitude of fig. 5.2-B(b) the two voltages will add during the second half, resulting in the old 100% modulation picture. So the circuit produces AM without carrier, or a DSB signal.

A word or two about circuit efficiency is now in order. Since we are screen modulating, the efficiency will vary from zero at no audio drive to normal class-C efficiency at audio peaks. If an analysis is made, the efficiency (based on average r-f power out to dc power in) will be:

$$\frac{\pi}{4}\,\eta M \times 100\%$$

for sine wave audio where ηm is the efficiency at the audio peaks which runs about 0.8. The overall efficiency is theoretically about 60% with 50% the value usually obtained in practice. This may not sound too impressive but let's look a bit further. Note that the efficiency expression involves $\pi/4$ and the normal class-C efficiency as a product. In a normal AM transmitter $\pi/4$ is the theoretical efficiency of the class-B modulators and ηm of course is the efficiency of the class-C final. Thus the circuit of fig. 5.2-A will produce r-f sidebands with the same efficiency as a conventional high level modulated AM transmitter. The reason the DSB final is so much simpler than a normal AM rig is that we don't transmit the carrier.

The peak power outputs which can be obtained from a given pair of tubes in this service may be estimated by taking the carrier output given in the handbook for one tube in class-C telephone service and multiplying by four. You can do at least this good and probably better. For example, if a pair of 6146 tubes is to be used, we find in the handbook that one tube will give 52 watts of carrier output in class C telephony service at 600 plate volts and 150 screen volts. If we set the high voltage at 1200 volts and run the screens to 300 volts on audio peaks we will get 4 x 52 or 208 peak watts output. This you know you can do because the voltages and powers quoted are those which exist in class-C telephony service during modulation peaks.

Without getting into too much circuit detail or DSB-linear amplifier power comparisons, this much is clear: the class-C amplifier with its ability to put out large amounts of peak power is ideally suited for voice service. The average voice sideband power produced by a pair of tubes in DSB service will easily match the average voice sideband power produced by the same tubes in SSB linear amplifier service.

The above power discussion actually underplays an important advantage of DSB over SSB. In DSB or standard AM systems voice clipping and filtering, if properly done, can increase significantly the average sideband power output of a given transmitter. Such tricks cannot be used in SSB since a flat-topped wave is deadly to an SSB system. (Such a waveform results in a very high peak-to-average power ratio for the SSB signal). Do not confuse peak clipping with the peak limiting or audio AGC tricks that are sometimes used in SSB designs. These are defensive measures which in effect permit the audio peaks to fully load but not overload the linear r-f amplifier. The average power gain of DSB using a good clipper-filter over SSB can be considerable.

The r-f excitation in DSB service is not all critical. Adjust for normal phone drive and leave it at that. This is one reason why screen modulation of tetrodes is to be preferred over control grid modulation of triodes. You can use triodes, of course, but you will have to watch the ratio of audio voltage to r-f voltage. This is true for any grid modulated balanced modulator that uses triodes. It should also be mentioned that the circuit shown in fig. 5.2-A is self neutralizing since the grid plate capacitance of the non-conducting tube is just right to neutralize the conducting tube.

Receiving Double Sideband

Contrary to what you might hear or read, DSB will sound no different than SSB when heard on a selective receiver. As proof of this, consider the 75A4 or other receivers employing selective band pass filters. Any phone signal when properly tuned in on such a receiver, is automatically converted to SSB. The suppressed carrier frequency is positioned approximately 20 db down the skirt response. The received sideband will fall within the band-pass frequency of the receiver, and the other sideband (if present) will be rejected. The communications receiver simply is not "smart enough" to know that the original signal was DSB, rather than SSB. By the time it reaches the detector it is SSB.

However, it may be more difficult to *tune* DSB, than SSB, for two reasons. First, the products from the "other" sideband will be present (and audible) until the suppressed carrier frequency is properly positioned on the response curve. It does take a bit of a knack to tune the DSB signal, but so does SSB for that matter, and obtaining the knack should be no great obstacle.

The second reason is far more important. When transmitting DSB, low frequency intermodulation distortion will be very noticeable, unless the "lows" are suppressed in the transmitter. The low frequency components in the voice will beat with the inserted carrier and produce an indescribable distortion component. This effect is particularly objectionable when the phase of the inserted carrier is *close* to the phase of the suppressed carrier. The solution is simple, however. It is necessary to "roll-off" the low frequency end of the audio response curve in the modulation section of the transmitter. This can be accomplished by using small coupling capacitors (.001 mfd.) between stages, and small cathode by-pass capacitors (4 mfd.). The additional use of a Stancor C-2340 band-pass filter (300- 3000 cycles) will also help to minimize this intermodulation distortion. It should be obvious from the proceeding discussion that it is not necessary to receive DSB with phase lock equipment. The same stability required for SSB will also suffice for DSB. A double sideband receiving adapter, employing phase synchronous detection will make maximum use of the system, however. Such a unit is described later in this chapter.

Modulation Transformer

The only component that warrants special mention is the modulation transformer. Many hams have evidenced difficulty in obtaining suitable units, with the necessary step-up ratio. A useful modulation transformer must have at least a 3 to 1 step-up ratio (to provide the necessary screen grid voltage) and the secondary must be center-tapped. In addition, the unit should be large enough so that excessive IR losses and core saturation are not encountered. Experimenters have been very successful using old power transformers! Any device delivering about 300 volts rms each side of the center tap (assuming a 117 volt primary, of course) will work. A war surplus 400 cycle transformer with these ratings would be a "gem". These 400 cycle transformers usually have good frequency response up to several thousand cycles, but poor response below 300 cycles. Thus the transformer response will complement the audio response ideally.

Several commercial transformers have proven themselves to be "naturals" for this DSB modulator application. For rigs up to, and including the 6146 class, the Thordarson-Meissner 20D79 is ideal. This transformer has a 5.2:1 winding ratio and is only 2-7/8 x 2-1/16 x 2-3/16 inches. Originally, this unit was intended to drive push-pull class B grids, hence, it is actually a step-down transformer. But for our application, it works very nicely by reversing the windings, that is: the green lead goes to the modulator plate, the black wire to B plus, the red center-tap is grounded, and the brown and blue leads are connected to the screens of the DSB final. A type 6AQ5 tube will provide more than ample modulation with this transformer.

A larger transformer, for use with kilowatt finals, is made by Stancor and carries the part number A-4762. Again, this unit is intended for class B service (5:1 step down ratio) and must be "reverse connected". A pair of 6L6's with this transformer, will modulate a pair of 4-250's well into the illegal region.

5.3 A DOUBLE SIDEBAND TRANSCEIVER, FROM A COMMAND TRANSMITTER

This project is an illustration of how simple a DSB transmitter can be, even on the high frequency bands. It consists of a Command Set transmitter which has been modified to generate a DSB signal and is self contained, except for the power supply. In addition, two accessory units will allow the device to function as a transceiver, when used in conjunction with a car radio. The units are constructed separately

Parts List for Fig. 5.3A

C1—100 mmfd. silver mica
C2—Original padding capacitor—C60
C3—Original tuning capacitor—C63
C4—200 mmfd. (original component—C73)
C5—.006 mfd. (original component—C61)
C6—.002 mfd. (original component—C74)
C7—20 mmfd. silver mica
C8—200 mmfd. silver mica
C9—.001 mfd. disc ceramic
C10—50 mmfd. air padder capacitor (Johnson 50K10, 158-3)
C11, C12, C13, C14—.005 mfd. 600 volt disc ceramic
C15—Dual section 100 mmfd. variable capacitor (Johnson 100LD-15)
C16—500 mmfd., 10Kv. (Centralab TV5-501)
C17—20 mfd., 450 volt electrolytic
C18, C19, C20—10 mfd., 25 volt electrolytic
J1—Motorola style antenna connector
J2—UHF style antenna connector (Amphenol)
J3—Three pin microphone connector
L1—Original oscillator coil (T53)
L2—38 turns, #30 wire, closewound on ¼" form. Link is 2 turns, #22 plastic covered wire. (Miller #4500)
L3—11 turns, #16 wire close wound on ½" form. Link is 5 turns, #22 plastic covered wire. (National XR-50 form)
L4—See Fig. 5.3-C
M1—0-3 ampere R.F. antenna current meter (Arrow Sales, Inc.)
M2—0-300 milliampere meter (Arrow Sales, Inc.)
R1—51K, ½ watt (original component—R72)
R2—47K, 2 watt
R3—330K, ½ watt
R4—100K potentiometer, linear taper (Centralab Radiohm—B40)
R5—10K, 1 watt
R6, R7—47 ohm, ½ watt
R8, R9—100 ohm, 2 watt
R10—100 ohm, ½ watt
R11—100 ohm, wire wound potentiometer (Centralab WW101)
RFC1, RFC2—4 turns, #22 spaced to fill length of R8, R9.
RFC3—17 mh. choke (original component L53) Note: some Command Sets do not use this component. The Miller 5737 is a suitable replacement.
RY1—Original keying relay, K53 (see text)
S1—SPST toggle switch
T1—Modulation transformer (Thordarson 70D29)
T2—Modulation transformer (Stancor A-4742 or Triad A-1X)
V1—12SK7
V2—12A6
V5—12V6
V3, V4—6146 (or parallel connected 6883)

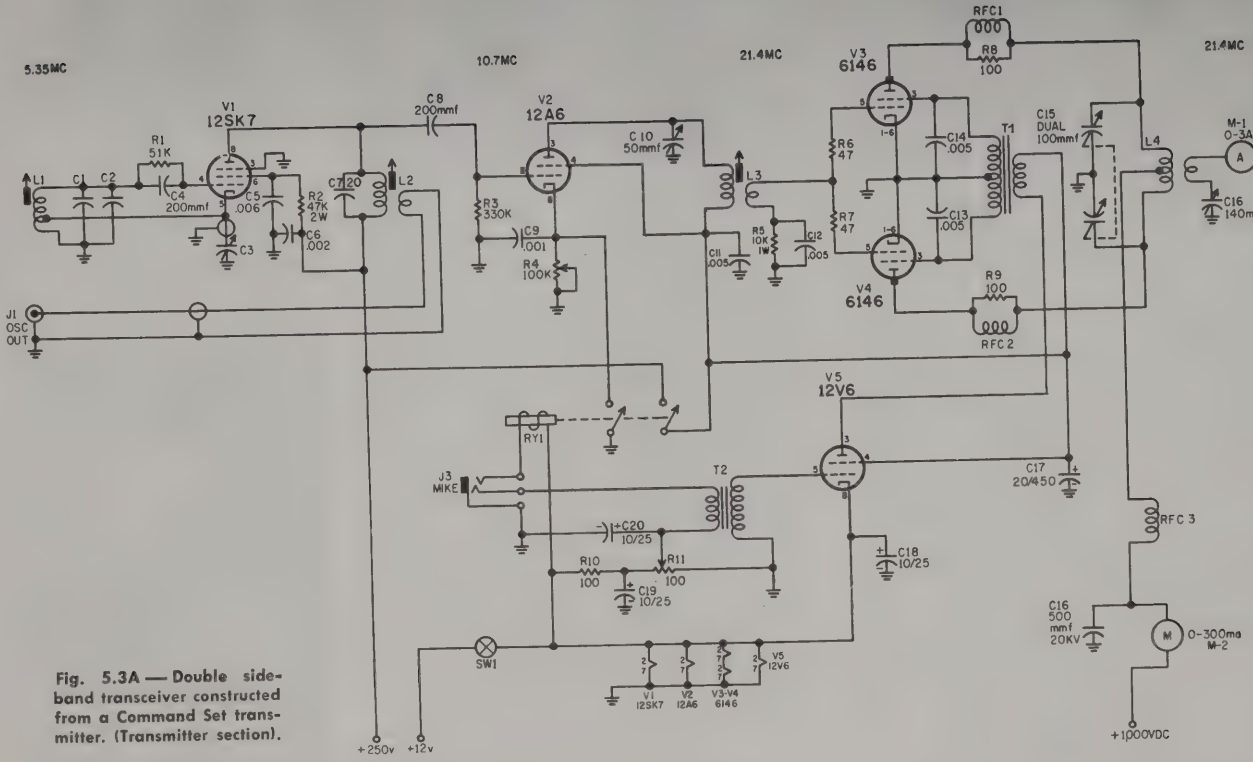


Fig. 5.3A — Double sideband transceiver constructed from a Command Set transmitter. (Transmitter section).

so that this feature can be omitted, if desired. The Command Set transmitter provides all the signals necessary for carrier re-insertion and provides a suitable local oscillator signal for the converter. The complete installation is designed so that the local oscillator signal will track with the transmitter. In effect, the transmitter oscillator is used to tune the receiver. The output of the converter is on 1600 kc and the car radio is adjusted to this frequency. All other adjustments are independent of the car radio, with the exception of the volume control. Since carrier re-insertion is accomplished at the signal frequency, it is not necessary to modify the car radio to install a BFO.

The foundation for the transmitter is the old faithful BC-457 Command Set transmitter. Actually, *any* model can be used so long as the frequency can be padded to cover the correct oscillator range (5.3- 5.4 mc). This particular unit was purchased from Arrow Sales, Inc. for $5.95, and appeared to be brand new. Any price between six and nine dollars for a near new unit seems quite reasonable. You sure couldn't build it for that! The original 1625's were removed and 6146's installed in their place. The original tubes can be used with only a slight loss of power and at the expense of somewhat more modulation power. The modulator shown (V5-12V6 in fig. 5.3-A) will provide ample modulation for either set of tubes. With 1000 volts applied to the plates of the 6146's the power output is approximately 110 watts peak.

Transmitter Section

Referring to fig. 5.3-A, you will note that the oscillator stage has been modified to use a 12SK7 tube as an electron coupled oscillator. Coil L1, in the control grid circuit determines the oscillator frequency and tunes between 5.3 and 5.37 mc. A tank coil in the plate circuit is tuned to the 2nd harmonic and drives the 12A6 doubler (V2). The 10.7 mc energy is also link coupled to the receiving converter through a jack on the rear apron. This signal is subsequently modified to provide the local oscillator signal. The 12A6 (V2) doubles the input signal to 21.4 mc which in turn drives the final amplifier. A potentiometer in the cathode of V2 is used to control the amount of carrier injection when receiving. Although the tube is not conducting during stand-by, the action of the control is very effective. When the push-to-talk mike button is depressed, relay RY-1 trips and one set of contacts shorts out the carrier insertion control. The 21.4 mc carrier energy drives the 6146 grids in parallel and the output circuit is connected in push-pull. The screens are modulated in push-pull. This is just the opposite hook-up of that shown in fig. 5.2-A, but serves to illustrate that either configuration is equally effective. Connected either way, the stage produces a double sideband, suppressed carrier signal.

A 12V6 tube is employed as a modulator. Note that a "trick circuit" is used to bias the tube. The cathode of V5 is returned to plus 12 volts, whch means that the

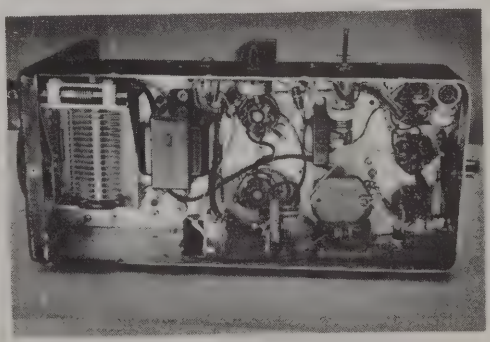
Transmitter portion of the DSB transceiver using surplus BC-458 (ARC-5) as the foundation. The meter on the left reads antenna current and the one on the right reads balanced modulator plate current. The knob between the meters adjusts the carrier insertion on receive, while the knob directly below resonates the final amplifier. The antenna loading adjustment can be reached through the hole directly below the antenna current meter. Access for the grid capacitor is through the tuning instruction plate.

cathode is more positive than the control grid (or the grid is more negative than the cathode) by 12 volts. This allows the voltage, that would normally be used for self bias, to be applied in the output circuit. Since the class A bias value for this tube is 12 volts, it works out nicely. If your automobile has a positive ground, it will be necessary to ground the cathode, lift the bottom end of T2 from ground, and connect it to the filament string. This will apply negative 12 volts to the grid through T2. Should you own a car with a 6 volt system, it will be necessary to use a 12 volt bias battery, or to insert a 270 ohm, 5 watt resistor between cathode and ground (self bias). Also, for 6 volt systems, the value of R10 should be reduced to 33 ohms, V1 should be a 6SK7, V2 and V5 should be 6V6's, and the filaments of V3 and V4 should be parallel connected. This transmitter can also be used on an ac supply, but it will be necessary to employ a microphone battery, and the modulator will have to be wired for self bias operation. The modulation level of the transmitter is controlled by R11, which regulates the amount of microphone current. If you are unable to obtain sufficient modulation, reduce the value of R10, but be careful not to exceed the microphone current rating. Note that the microphone voltage is also obtained from the car battery. The B supply is applied to the oscillator at all times for better stability and for carrier insertion. When relay RY-1 is energized, the B plus is applied to all the other stages. Although not shown, an extra set of contacts on the antenna relay is used to remove B plus from the converter (V1-V2, fig. 5.3-C) when transmitting.

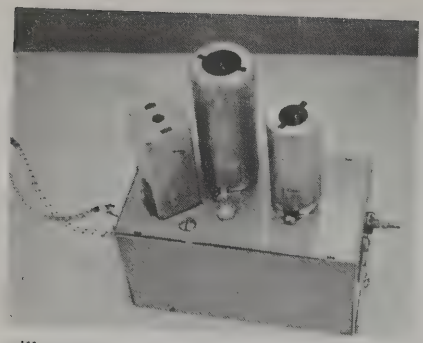
Two meters are used to check the operation of the transmitter. A 0 -300 ma meter indicates the plate current and the pointer just "taps the pin" in actual operation. Antenna current is displayed by a 0 -3 thermocouple r-f ammeter connected in series with the "hot lead" of the antenna link. In operation *everything* is adjusted for maximum antenna current (with modulation), as indicated by maximum on this meter. When this has been done, the proper loading conditions for DSB operation are automatically established. The transmitter power requirements are: high voltage- 1000 volts, 300 ma short duty cycle, low voltage- 250 volts, 70 ma continuous duty cycle.

Receiving Section

The circuitry associated with V1 and V2 is (fig. 5.3-C) standard for a high frequency converter, except that it contains no local oscillator. Tube V3 (a 12SA7) provides this signal. The 10.7 mc. energy is fed into the 12SA7 through a short length of coax which plugs into the receptacle on the rear of the transmitter. Coil L4 is resonant at the 2nd harmonic, or 21.4 mc and it is this frequency that is actually applied to V3, the mixer. The 21.4 signal mixes with a 1600 kc local oscillator and the beat sum frequency (23.05 mc) appears in the output circuit (T2). Actually, since the transmitter input signal tunes between 10.6 and 10.74 mc, this



The microphone jack, volume control, and final grid tank are mounted on the side apron of the BC-458 (ARC-5). The front capacitor is used to tune the vfo and is connected to the oscillator coil through a short length of coax cable. The capacitor in the lower center area resonates the grid tank circuit.

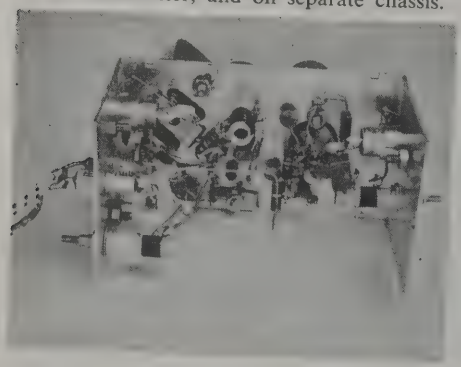


This is the 15 meter converter section of the DSB Transceiver. The grid coil for the r-f amplifier can be seen at the right, and the plate coil is between the amplifier and mixer. The 1600 kc output transformer is at the left.

will produce a beat frequency which can be tuned between 22.8 and 23.05 mc. It is this frequency that appears at connector J4 and is fed to the local oscillator input jack (J3 at V2). This beats with the incoming station and produces the 1600 kc i-f signal that is fed to the car radio through connector J2. A type 6CB6 is used as an r-f amplifier and is resonated by coils L1 and L2. Transformer T1 is tuned to 1600 kc, T2 is staggered slightly to cover 22.8- 23.05 mc, and L5 is the 1600 kc oscillator coil. Coil L6 and capacitor C19 are resonant at 21.4 mc and are used to reject any fundamental frequency energy that might get through the mixer tube (V3) and transformer T2. At the other end of this cable, coil L3 is series resonant at the center of the local oscillator frequency range (22.95 mc) and provides additional rejection of the fundamental frequency. The car radio itself provides sideband selection. If the radio is tuned to slightly less than 1600 kc. the upper sideband will be received. If the radio is tuned to slightly higher than 1600 kc. the lower sideband will be received. This will not affect the transmitted signal however.

Precautions

As you can well imagine, it is a bit of a trick to generate enough signal for oscillator injection, without "overpowering" the input to the converter. If excessive signal is allowed to leak outside the transmitter case there will be so much carrier re-insertion that the AVC will practically disable the car radio. When testing the unit, all the cases should be "buttoned up" tight (covers in place, etc.). To this end, it was also necessary to add L6 and C19 as a tuned trap for the signal frequency (21.4 mc.) in the local oscillator lead. Short cables must be used between J1 (on the transmitter chassis) and PL1 on the converter chassis, and between J3 and J4 on the converter chassis. It is also necessary to mount the converter-mixer unit external to the transmitter, and on separate chassis. Unless the converter is separate from



Underside view of the 15 meter converter section of the DSB Transceiver. The coil in the lower left corner matches the coax from the local oscillator to the mixer grid. Note the use of feed-through capacitors.

This is the local oscillator mixer. It converts the Command Set transmitter vfo to a frequency suitable for injection in the converter. With this system, the receiver and transmitter "track", or operate on exactly the same frequency. The buttons on the rear of the chassis are feed-through capacitors.

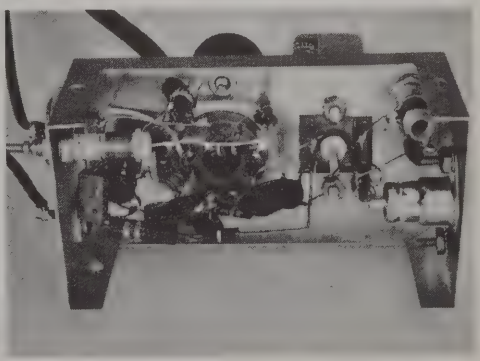
the transmitter it will be very difficult to keep the injection voltage low for weak signals. And unless the converter and mixer are in separate boxes, spurious signals will be received. With the arrangement shown, extra carrier injection will be needed on all but the weakest stations. Additional carrier injection is available by turning up potentiometer R4, located on the front panel of the transmitter.

The car radio must be reasonably selective for good sideband reception. However, since this is seldom the case, it might be wise to incorporate a selectivity device between the output of the converter and the input to the car radio. An excellent device for this purpose is described by Denny Moore, W6MHP, in the February 1955 issue of QST Magazine. It is a passive inductance filter and uses no vacuum tubes, therefore requires no connection to the power circuits.

Another precaution is in regard to oscillator stability. There should be no more than a few cycles shift of the oscillator frequency between transmit and receive. If there is, and you are working another transceiver, the two stations can walk right out of the band!

Construction

Oscillator section: The 12SK7 oscillator (V1) is mounted in the same socket as the original, but the circuitry associated with this stage is re-wired to conform to fig. 5.3-A. The cathode tap for the oscillator stage is placed four turns up from the ground end of the coil. The front capacitor is used for oscillator tuning and is connected, through a short length of shielded wire, to the cathode connection. This lowers the effectiveness of the capacitor and decreases the tuning range. By proper juggling of the slug and padder associated with L1, the band (from 21.2 to 21.45) can be spread over the entire dial. Coil L2 is mounted directly above the socket for V1 and the output jack (J1) is just to the rear of the coil. The 12A6 is



Underside view of the local oscillator mixer. The coil at the left steps up the impedance of the coax carrying the vfo signal, while the coil at right is a trap for the vfo energy.

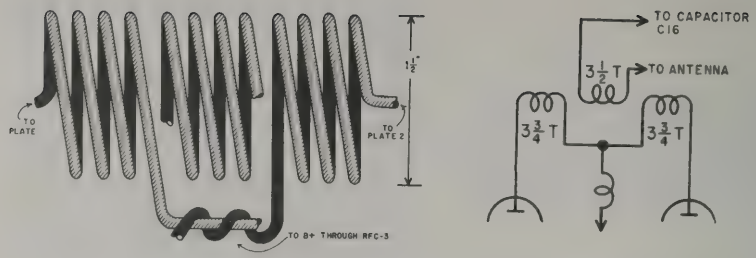


Fig. 5.3B—Winding details for the plate tank inductance. The coil stock is Air-Dux 1210T.

located in the socket formerly occupied by the calibration crystal. The cathode of this tube is connected to the relay contact through a length of shielded cable. Note that one of the relay contacts is permanently grounded and it is the associated contact that is used to short out the cathode potentiometer. Incidentally, this original relay may be used on 12 volts if the tension is reduced slightly. For 6 volt systems, the relay will have to be replaced. The plate of V2 (12A6) is connected to coil L3 with a length of heavy bus-bar wire, and the wire continues on over to C10. A link on L3 is used to match the impedance of the 6146 grids. A tie point is mounted near L3 and is used to secure the grid resistor R5 (10K) and its associated by-pass capacitor (C12- .005 mfd.).

The 6146's are mounted on a new plate that covers the 1625 holes. The modulation transformer is mounted on the underside of the chassis between the 6146's and the oscillator tuning capacitor. On the top side of the chassis will be found RFC1 & 2, C15, C16, L4, RFC3, and C16. The antenna connector is mounted directly above the antenna current meter. The carrier re-insertion potentiometer is located between the two meters. The knob for the final tuning capacitor is directly above the oscillator frequency dial. Capacitor C16 is used to tune out the link reactance (and affect loading) and is adjusted through an access hole directly below the antenna current meter. A small hole to the left of the oscillator dial lines up with the drive capacitor (C10) and can be adjusted with a long screwdriver. Originally a shaft was brought out to the front but it was found that the grid circuit seldom required tuning.

The modulator tube is wired in the socket position formerly used for the tuning eye. The by-pass components associated with this tube are mounted on the rear apron, adjacent to the socket. The microphone transformer (T2) is located below the oscillator compartment and directly above the slug tuned coil, L3. The microphone jack and modulation control are mounted on the side of the chassis in line with the 6146's.

Converter Mixer Section

The converter unit is mounted on an LMB #107 chassis box. The antenna input jack (J1) and grid coil (L1) are mounted on one end. Tubes V1 and V2, and transformer T1 are mounted on the top, and staggered slightly. Coil L2 (control grid, V2) is mounted on the top of the chassis beside mixer tube, V2. On the other end of this chassis will be found J3, with the adjustment for L3 directly below it. J2 (output to the receiver) is also located on the rear chassis apron, along with the feedthrough capacitors for filament and B plus.

The cable emerging from the mixer chassis (12SA7-V3) goes to the oscillator output on the transmitter chassis. The two feedthrough capacitors for filament and B plus are located on one end of the chassis along with the slug adjustment for L4.

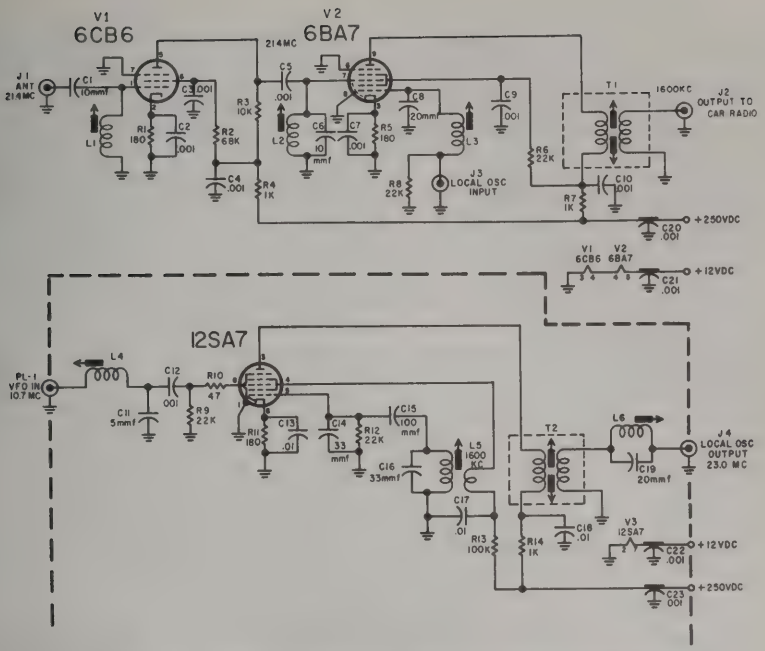


Fig. 5.3C—Double sideband transceiver (converter-mixer section)

The miniature oscillator coil slug protrudes through the chassis directly in front of the 12SA7 tube. Transformer T2 is mounted in line with the 12SA7, along with coil L6. The local oscillator output jack (J4) is mounted on the rear apron of the chassis box. The mixer section is constructed on a L. M. Bender chassis box measuring 4 x 2⅛ x 1⅝ inches (TF-774).

Adjusting the Units

The transmitter tune-up is simplicity itself and need not consume much time. Adjust the tuning range of coil L1 so that its 4th harmonic just covers 21.2 to 21.45 mc by varying the padder capacitor and slug associated with coil L1. It is permissible to pull a few plates off tuning capacitor C3, if necessary. Use extreme care, for the rotor is mounted in ball-bearing arrangement, and is easily ruined. When you are satisfied with the tuning range, adjust coils L2 and L3 for maximum grid voltage as measured across resistor R5. The voltage should be between 30 and 60

Parts List for the Double sideband transceiver (converter-mixer section)

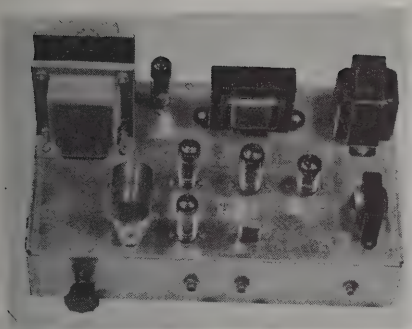
C1, C6—10 mmfd. silver mica
C2, C3, C4, C5, C7, C9, C10, C12—.001 mfd. disc ceramic
C8, C19—20 mmfd. disc ceramic
C11—5 mmfd. disc ceramic
C13, C17, C18—.01 mfd. disc ceramic
C14, C16—33 mmfd silver mica
C15—100 mmfd. silver mica
C20, C21, C22, C23—.001 mfd., 60 volt chassis feedthrough capacitor
J1, J2, J3, 14—Motorola type antenna connector
L1—23 turns, #30 close-wound on ¼" form (Miller 4505)
L2—20 turns, #26 close-wound on ¼" form (Miller 4500)
L3, L4—18 turns, #30 close-wound on ¼" form (Miller 4504)
L5—Broadcast band oscillator coil (Miller 70-OSC)
L6—23 turns, #30 close-wound on ¼" form (Miller 4505)
PL1—Motorola style antenna plug
R1, R5, R11—180 ohms, ½ watt
R2—68K, ½ watt
R3—10K, ½ watt
R4, R7, R14—1K, ½ watt
R6—22K, 1 watt
R8, R9, R12—22K, ½ watt
R10—47 ohms, ½ watt
R13—100K, 1 watt
T1—1500 kc. i-f transformer (Miller 13-W1)
T2—21.25 mc. i-f transformer (Miller 6190-TV)
V1—6CB6
V2—6BA7
V3—12SA7

volts, negative. Next, apply plate voltage and modulate the transmitter into a dummy load. Adjust C15 and C16 for maximum power output as indicated by maximum r-f current. The transmitter is now ready to operate.

The converter-mixer unit will take a little more time, since there are many more adjustments. First, insert a two turn coupling link in J3 (local oscillator input) and apply energy from a grid dipper tuned near 23 mc. You should be able to hear 15 meter Amateur stations. Tune in a station and peak L1, L2 and T1 for maximum signal strength. Next, energize the mixer chassis and couple a small amount of signal from L5 to the output jack, J2. This should put a beat note on any AM station that you are listening to. Now, remove the grid dipper and the link from J3, and connect the cable from J3 to J4. Disable the 12SA7 oscillator by shorting pin 5 to chassis ground. Connect the link and grid dipper to PL-1, with the GDO still tuned to approximately 23 mc and tune in an AM station. Peak T2 for maximum signal strength. At this point, it is a good idea to recheck the setting of L1, L2, and T1. With covers and bottoms on all units, apply the 10.7 mc oscillator signal to V3, through PL-1. Tune the car radio slightly off 1600 kc to avoid overload at this point in the adjustment process. Tune in an AM station by adjusting the oscillator dial on the transmitter. Now, tune coils L3, and L4 for maximum signal strength. (Note: these adjustments will be rather broad, but a maximum should be easy to locate.) Now, retune back to 1600 kc and a strong unmodulated signal should be heard (even with the carrier injection knob turned off). This is the fundamental signal from the transmitter. Adjust coil L6 for *minimum* strength of this signal. If this adjustment does not seem to affect the strength, the signal is probably getting into the converter chassis from another source. At this point, connect a temporary by-pass to all power leads entering or leaving the three chassis, until the signal is as weak as you can get it. Remember, the less the carrier injection, the weaker the station can be, for Q5 copy. At this point, you should be able to tune in SSB or DSB stations. Remember to set the car radio dial to slightly less than 1600 kc for upper sideband reception. As a "final alignment", peak L1, L2, T1, T2, L4, and L3 for maximum strength, but stagger them slightly so that the gain is fairly constant between 21.2 and 21.45 mc. Of course, if you are only interested in the "SSB section" of the band, you can ignore the other frequencies. If there is a strong local broadcast station near 1600 kc, the frequency may be changed by varying the setting of L5 to any convenient frequency. Stay above 1000 kc though, or the image problem could be severe. It should be noted that you can copy AM stations very nicely by tuning off 1600 kc (to eliminate the beat note), but then, who wants to? Another point that should be stressed: you can use this system for any of the ham bands up to 10 meters. The same theory will work. In other words, you apply a signal to the signal mixer at the frequency you are receiving. This beats with an oscillator that is tuned to the frequency the car radio is set-up for, and this produces a tunable local oscillator signal. When this is mixed with the energy from the incoming station, we convert *back down* to 1600 kc and it is this signal that is applied to the car radio. For 80, 40, and 20 meter DSB transceivers, the 12A6 doubler could be eliminated. It should also be pointed out that the DSB transmitter does not have to be used with the "tracking converter." Simply eliminate J1 and the link on L2 for regular operation.

5.4 THE DOUBLE SIDEBAND JUNIOR

The author is grateful to the General Electric Company and Lighthouse Larry, for permission to reproduce K2GZT's popular DSB rig. This unit was originally featured in *GE HAM NEWS* (Mar-Apr 58, Vol. 13- No. 2). The DSB Jr. (fig. 5.4-A) output stage consists of two type 6AQ5 pentode tubes (V2 and V3)



Top view of the double sideband transmitter showing the locations of major parts on the chassis deck. Check to see that sufficient space is provided for components which differ in size and shape from those listed. The audio filter inductor (12) and the microphone transformer (T3) should be oriented in the positions shown to prevent inductive hum pickup from the power transformer.

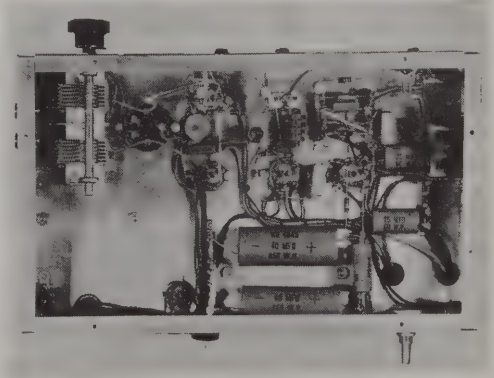
with the control grids in parallel, and the screen grids and plates in push-pull. This balanced modulator circuit was chosen because a compact receiving type two-section variable capacitor (C1) can be used in the push-pull plate tank circuit. The r-f output is link coupled from the center of the plate tank coil (L2).

The grids are driven by a crystal controlled oscillator, one half of a 12BH7 twin triode tube (Vla). The other half (Vlb) is the audio modulator stage. The r-f output stage is screen modulated with the push-pull audio signal, transformer coupled from the modulator stage. The transformer specified for T2 is connected backwards (primary to the screen grids of V2 and V3; secondary to plate of Vlb). The r-f carrier signal applied in parallel to the control grids of the 6AQ5 tubes is cancelled out in the push-pull plate circuit.

With no modulation the plate current in both final tubes will be low because of the low screen voltage. If a sinusoidal audio tone is assumed as the modulating signal, one screen is driven positive during the first half-cycle and the other is driven negative. The 6AQ5 having positive screen grid conducts and an r-f current is supplied to the load by that tube. During the next half of the audio cycle, the other tube supplies r-f power to the load and the first tube rests. Note that only one tube is working at any one time, except when there is no audio; then both tubes rest. Neutralization is no problem, as the balanced modulator circuit is self-neutralizing.

A positive bias for the 6AQ5 screen grids—about 13 volts—is developed across the 2000 ohm resistor in series with the cathode-to-chassis connection for the modulator tube (Vlb). Current for operating a carbon microphone is supplied through the 1500 ohm resistor.

The two audio voltage amplifier stages employ a 12AU7 twin triode (V4). The first stage is driven by a single button carbon microphone through a matching trans-



Bottom view of the chassis, showing placement of smaller parts on the tube sockets and terminal strips. Power wiring is run in corners and across the center of the chassis. Wires carrying audio and r-f voltages should be made as short as possible.

former (T3). The first audio stage drives a shunt type diode clipper circuit which clips both positive and negative audio signal peaks. The clipping level is adjusted by varying the positive bias on the clipping diodes, D1 and D2. This bias is obtained from a 1000 ohm potentiometer in series with the cathode-to-chassis circuit of the second audio amplifier stage (V4b).

A simple pi-section audio filter (C2, C3 and L3) following the clipper suppresses the audio harmonics ("splatter") generated in the clipping process. The second audio stage then drives the modulator (V1b).

Push-to-talk operation of the transmitter is obtained simply by grounding the cathode of the crystal oscillator tube (V1a) through a single pole, single throw, normally open push-button switch of the type found on most single button carbon microphones (war surplus T-17, or Electro-Voice Model 210-KK). If the push-to-talk feature is not desired, substitute a two conductor phone jack for the three conductor jack (J3) shown in the schematic diagram.

Additional audio amplification will be required if a low-output crystal, ceramic, or dynamic microphone is used with the transmitter, in place of the carbon microphone. This extra gain can be obtained with a 12AX7 twin triode tube in a two-stage audio preamplifier. The circuit for this amplifier, which will deliver a voltage gain in excess of 1000, is shown in fig. 5.4-A. The arm on the 250,000 ohm gain control at the output of the second stage (V6b) feels directly into the grid of V4a. The transformer (T3) and carbon microphone voltage circuit can thus be eliminated. However, it will be necessary to ground the end of the 1500 ohm resistor that originally was connected to the microphone. In addition, the cold end of RFC-1 will have to be grounded so that the oscillator tube current path is completed.

The transmitter may be constructed with the high voltage power supply shown in the main schematic diagram. Any separate power supply capable of delivering 400 volts at 70 ma may be used instead. A lower plate supply voltage will result in reduced r-f power output, however. It is possible to use a plate supply delivering 500 volts dc. Although this exceeds the plate voltage ratings for the 6AQ5's, the plate dissipation will not be exceeded, and the tubes will not be damaged. No matter what type of supply is used in conjunction with this transmitter, it is recommended that a choke input filter system be used, for best regulation.

The 6AQ5 plate tank circuit, C1 and L2, should tune to resonance at 3.8 mc with C1 near maximum capacitance. If it will not tune this low in frequency, add a small padding capacitor—a 10 mmfd, 2000 volt working mica is suitable—across the ends of L2 on the plug-in coil base.

Mechanical Details

The transmitter was constructed on a 7 x 12 x 3 inch aluminum chassis

Parts List for Fig. 5.4A

C1—Two section variable, 7- 100 mmfd. per section (Hammarlund MCD-100S)
C2—500 mmfd., 500 volt mica
C3—300 mmfd., 500 volt mica
C4, C5, C6—25 mfd., 50 volt electrolytic
C7, C8 — 40 mfd., 450 volt electrolytic
C9—16 mfd., 450 volt electrolytic
D1, D2 — 1N63 germanium diodes (GE)
J1, J2 — Two conductor, closed circuit phone jack
J3—Three conductor, open circuit phone jack
L1—15 uh., 50 turns, #28 enam. wire scramble wound on ¼" of a ⅜" dia. slug tuned form (Miller #4400)
L2—44 uh., 48 turns, #22 wire, 1½" long, 1¼" dia., with 3 turn link at center (B & W 80JVL)
L3—6 hy., 40 ma. 300 ohm filter choke (UTC- R-55 or equiv.)
L4—14 hy., 100 ma. 450 ohm filter choke (UTC R-19 or equiv.)
R1—1K, 2 watt potentiometer, linear taper
R2—3K, 5 watt wire wound resistor
R3 — 250K, potentiometer, audio taper (Centralab Radiohm B-51)
RFC1—2.5 mh. r.f. choke
S1 — Single pole, single throw toggle switch
T1 — Power transformer — 880 volts c.t., 75 ma. four 6.3 volt filament windings (Triad R-70A) If 6X4 is replaced with a 5Y3, the Chicago PCR-85 may be used
T2 — Driver transformer, 5.2:1 turns ratio, reverse connected per test (Thordarson 20D79)
T3 — Carbon mike to grid transformer 31.4:1 turns ratio (Triad A-1X)
V1—12BH7 tube (GE)
V2, V3 — 6AQ5 tube (GE types 6005 or 6669 are also suitable)
V4—12AU7 tube
V5—6X4 tube
V6 — 12AX7 tube (optional audio amplifier)

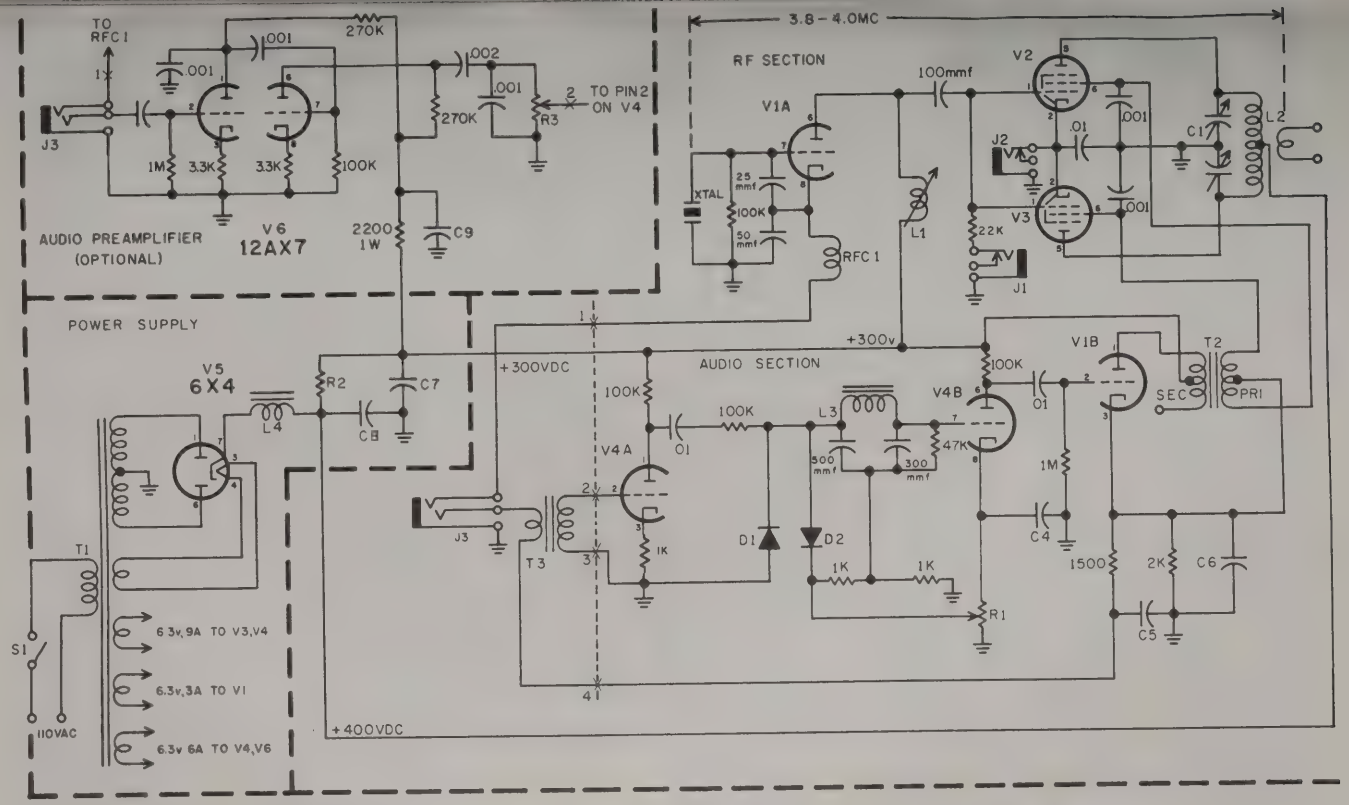
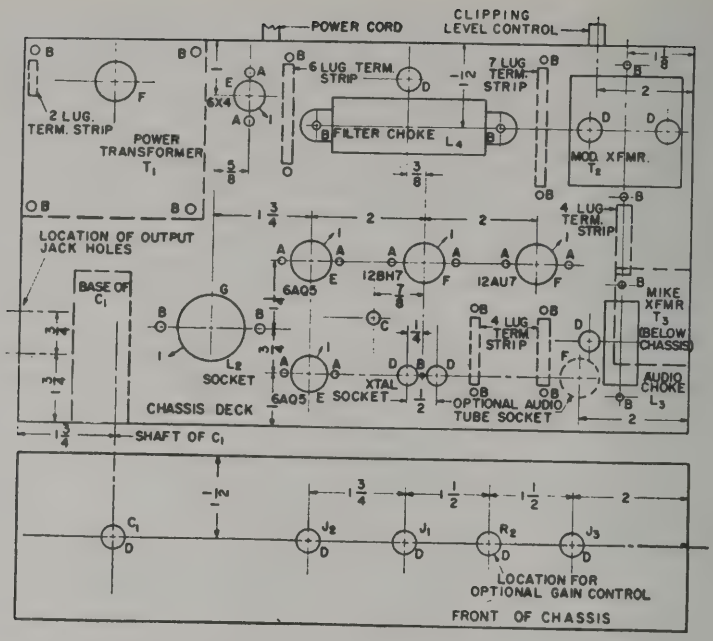


Fig. 5.4A—Diagram for the complete 20 watt double sideband transmitter. The optional audio preamplifier appears in the upper left hand corner. Capacitances given in whole numbers are mica, 500 volts working and those in decimals are disc ceramic, 500 volts working. Resistors are ½ watt unless otherwise specified.



DRILLING LEGEND

"A" drill—No. 32 for miniature tube socket hardware.

"B" drill—No. 26 for fastening terminal strips and larger components.

"C" drill—$\frac{9}{32}$ of an inch in diameter for L_1.

"D" drill—⅜ of an inch in diameter for controls, grommets, etc.

"E" socket punch—⅝ of an inch in diameter for 7-pin miniature tubes.

"F" socket punch—¾ of an inch in diameter 9-pin miniature tubes and grommet under T_1.

"G" socket punch—1 ¼ inches in diameter for L_2.

Fig. 5.4B—Chassis deck and front panel drilling diagram for the double sideband transmitter. Dimensions are shown from the edges of a 7 x 12 x 3 inch deep chassis. Tube sockets should be mounted with pin 1 in the position indicated at each socket hole. The socket for the optional audio preamplifier tube (V6) and gain control (R3) are located as shown.

(Bud AC-408). A smaller chassis, or utility box, will easily hold the r-f and audio components, especially if the power supply is constructed on a separate chassis. Of course, if a suitable high voltage supply is already available, utilize it instead.

The same relative locations for major parts, as shown in the drilling diagram, fig. 5.4-B, should be followed. If the audio preamplifier for low output microphones is to be included, the tube socket should be placed in the location indicated on this diagram. The matching transformer for a carbon microphone, T3, is then not required. The audio low-pass filter inductor, L2, should be mounted beneath the chassis in place of T3. The gain control between stages in the extra audio amplifier may be mounted midway between J1 and J3 on the front of the chassis, as indicated on the drawing.

Small holes for component fastening hardware should be located directly from the matching holes on each part; the drilling diagram simply indicates the presence, but not the precise location, of these holes. Rubber grommets should be placed in all chassis holes for transformer leads before these parts are assembled in the

locations shown in the top view photo.

The smaller parts beneath the chassis are fastened between tube socket lugs and lugs on other parts, or on lug-type terminal strips (Cinch-Jones 2000 series). Most of the audio clipper and low-pass filter components were assembled between two four-lug strips, as shown in the bottom view photo. Note that the tubular type electrolytic filter and cathode bypass capacitors fit neatly into unused portions of the chassis. Use of metal can type capacitors will require crowding of some components on the chassis deck.

All power and audio circuit wiring was run with #20 stranded, insulated hookup wire. Heavy tinned copper wire was used for the lead between the 6AQ5 control grid socket pins; also for connecting the 6AQ5 plate lugs to the socket for L2 and stators on C1. Small insulated banana jacks were mounted on one end of the chassis for antenna terminals, but a suitable chassis type coaxial cable connector may be substituted.

The audio preamplifier stage, which may be added to the transmitter at any time, was constructed on a turret type 9 pin miniature socket (Vector #8-N-9T). There is adequate room on this socket for all small parts, but the 16 mfd., 450 volt filter capacitor in the plate voltage decoupling filter should be placed in the corner behind T3.

ADJUSTMENT AND OPERATION

Once the transmitter has been completed, it should be tested on a dummy load consisting of a 15 watt, 115 volt incandescent lamp bulb. The bulb should light to nearly full brilliancy. The test procedure consists of the following steps:

1. Apply power and insert a crystal for the 3.8- 4.0 mc. phone band. Depress the microphone push-to-talk switch.

2. Adjust L1 to resonance while observing the final amplifier grid current on a milliammeter inserted at J1. A grid current of 3 to 4 milliamperes is required for proper operation.

3. Set R1 to its midpoint. Adjust L2 for closest coupling. Whistle into the microphone and adjust C1 for maximum output power or maximum brilliance of the dummy load lamp.

4. Observe the r-f output voltage with an oscilloscope. Either the bowtie or envelope presentation may be used. Whistle into the microphone. Successively adjust the output coupling and clipping level (R1) for maximum output voltage consistent with linearity.

5. Upon successful completion of testing with a dummy load, the transmitter may be connected to a transmitting antenna. The antenna should preferably be a low impedance tuned antenna, such as a dipole or folded dipole. If a long wire antenna is used, an antenna tuner should be used to transform the antenna impedance down to a value suitable for link coupling. When the transmitter is connected to the antenna, step 4 should be repeated to ensure that the output stage is properly adjusted and not overloading on positive audio peaks. The final amplifier cathode current may be metered at J2. The plate current will have a resting value of about 20 ma and will rise to about 40 ma with modulation.

Operating The DSB Jr. From A VFO

It was possible to feed the output from a Heathkit VFO directly into the crystal socket of the double sideband junior, on the 3.8 mc band, with good results. The connection may be made with a short length of RG-58/U coaxial cable. The triode oscillator circuit, acting as a buffer stage, did not go into oscillation. However, instability in this stage may be encountered with other types of vfo's. Make sure that the outer shield on the coaxial cable connects to the grounded terminal on the crystal socket.

Band	Crystal	L1	L2
7 mc.	7.204 - 7.296 mc.	7.5 uh., 40 turns, #28 enam. wire closewound, 5/8" long on 3/8" diameter slug tuned form (Miller #4400)	16 uh., B & W type JVL-40
14 mc.	14.204 - 14.296 mc.	3.7 uh., 27 turns, #28 enam. wire closewound, 3/8" long on 3/8" diameter slug tuned form (Miller #4400)	2.2 uh., B & W type JVL-15
21 mc.	21.254 - 21.446 mc.	2.2 uh., 18 turns, #24 enam. wire closewound, 3/8" long on 3/8" diameter slug tuned form (Miller #4400)	2.2 uh., B & W type JVL-15
28 mc.	28.504 - 29.696 mc.	1.2 uh., 10 turns, #24 enam. wire closewound, 5/16" long on 3/8" diameter slug tuned form (Miller #4400)	1.3 uh., B & W type JVL-10

Fig. 5.4C—A chart of coil and crystal values which will allow the DSB Junior to be used on other Amateur bands.

Using Other Power Transformers

The reader may be somewhat confused by the four filament windings on the Triad R-70A power transformer. This particular transformer is not a "junk box" item, but was specially selected for this job. The only commercial transformers available with more than 400 volts each side of the center tap are designed for use in regulated power supplies. This particular unit has extra filament windings for the regulator tubes. The 6X4 tube is usually considered a "bad actor" since it is prone to heater cathode shorts. However, in this circuit, such a short will probably not occur because of the isolated filament winding. Even if the heater shorts to the cathode, there will be no current flow, and no damage to the tube can occur. If you prefer to use a type 5Y3 rectifier tube, the Chicago PCR-85 is an excellent replacement. This unit has one 6 volt winding and one 5 volt winding. The B plus voltage will be identical but the current rating will be about 10 ma higher with the PCR-85.

Conclusions

No special effort has been made to achieve a high order of carrier suppression. However, laboratory measurements indicate 40 db of suppression in the original model. At least 30 db of carrier suppression should be obtained with reasonably symmetrical wiring in the r-f output circuit. In most cases, the audio hum and noise level will be about equal to the carrier level.

5.5 THE 75 METER SIDEWINDER

Many ham's first experience with sideband is on the 75 meter phone band and rigs designed for this frequency are very popular. The fact that the Sidewinder is designed for mobile or fixed station operation, and can be constructed from an old derelict Command Set transmitter, makes this a popular project. In addition, provision has been made for AM operation should this (ugh) be desired. A peak envelope power input of 100 watts can be obtained with the 600 volt power supply specified. However, as is customary in double sideband transmitters, the plate voltage can be run up as high as 1000 volts without exceeding the plate dissipation ratings.

Circuit Details

The command transmitter originally tuned 4.0 to 5.3 mc and this range can easily be padded down to cover the 75 meter phone band (as well as the cw band if desired) by adroit manipulation of the oscillator coil slug and padder capacitor. Using this range command rig has the added advantage of a higher "C" oscillator tank than would be the case if the 3.0 to 4.0 mc transmitter is used. This means better oscillator stability. The original 1626 oscillator is followed by a 6AK6 buffer It should be emphasized that some form of isolation between the oscillator and final is essential—otherwise the final will pull the oscillator frequency and you will have a novel system of FM plus double sideband less carrier which will not endear you to the fellows at the other end. Since ours was a 12 volt system, the 6AK6 heater was wired in series with the front panel #47 pilot lamp thereby conserving .945 watts of dc power. You think like this after years of mobiling.

The 6AK6 develops its drive across a low "Q" slug-tuned coil. A look at the schematic will show you how to get away from the nasty chore of center tapping this coil while still ending up with push-pull drive to the final grids. The small mica trimmer at the lower end of the coil compensates for the 6AK6 capacity across the

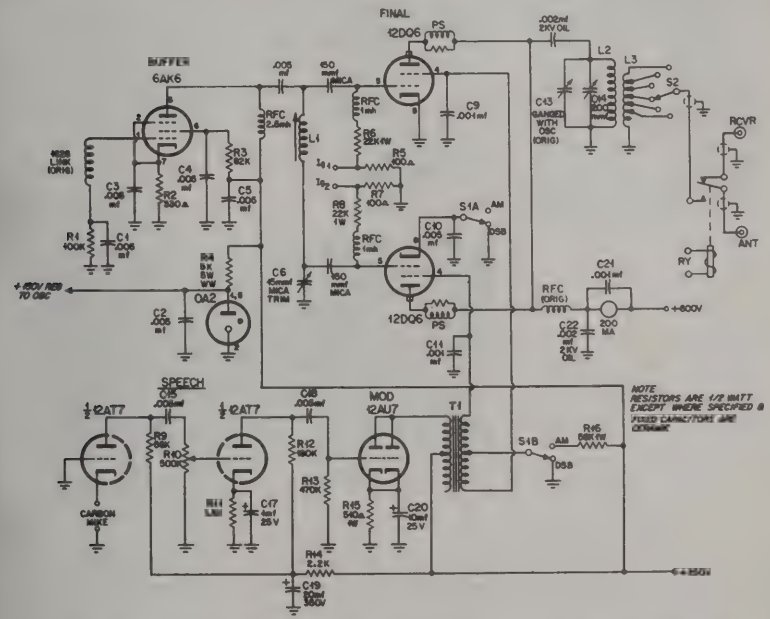


Fig. 5.5A—Schematic for the 75 meter "Sidewinder". The circuitry associated with the 1626 oscillator tube remains intact, and is not shown.

Parts List

L1—80 turns #28 enam. scramble-wound on ½" dia. slug-tuned coil form.

L2—30 turns #18 tinned, 1" dia. 2½" long. Air-wound with plastic ribs.

L3—10 turns #14 tinned, wound around bottom of L2. Space diameter of wire and cement to L2 with 1/16" concentric clearance from L3: Tap every turn.

S1—DPDT toggle switch.

S2—Ten position rotary switch.

T1—Interstage transformer. Turns step-up at least 1:2 modulator plate to screens. See text.

PS—5 turns #22 tinned wound on 100 ohm 1 watt resistor.

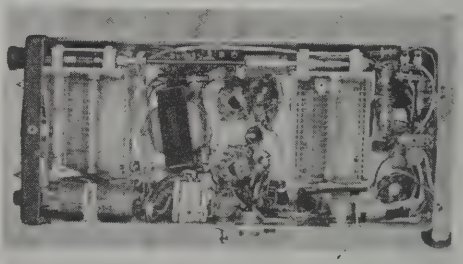
RY1—12 volt d.c. SPDT relay.

top side of this coil so you will end up with equal grid drive to each final tube. If you want to be different, you can drive the grids in parallel and operate the plates in push-pull and come out with the same results. In case you hadn't recognized it, this final is nothing more than our old friend, the push-pull doubler—except that in this application it is operated straight through. The result is carrier cancellation.

Separate grid rfc's and grid resistors are needed with this arrangement but this is desirable because we want to be able to look at the grid current for each final tube in the initial tune-up. This scheme of push-pull input can be considered self-balancing and should therefore give us better carrier cancellation, although this is apparently not a problem. At any rate, none of the stations worked to date have been able to find the carrier so it must be pretty well buried.

The 75 meter "Sidewinder". A self contained DSB transmitter constructed from a Command Set transmitter.

Top view of the 75 meter Command Set DSB transmitter. In the lower right corner is the 1626 oscillator and mounted from right to left are the 6AK6, 12AU7, and 12AT7.



Bottom view. Both of the original tuning capacitors are utilized so that the oscillator and amplifier track closely.

The final tubes are 12DQ6 TV sweep output bottles—big brother to the 12BQ6. Both these types have high perveance—that is, you can make them pull their load of plate current with comparatively low plate voltage. A second very desirable characteristic of this family of tubes is that the screen power requirements are relatively low. This means that the audio modulating power required for a given peak power output is correspondingly lower. Although the original 1625 tubes can be used, their higher screen power requirements may result in somewhat less peak power unless the audio section is beefed up. Although either the 12BQ6 or 12DQ6 can be used, we settled for the latter because of the higher plate dissipation rating (15 watts versus 11 watts) and slightly higher Gm.

The final tank is a conventional shunt-fed, single-ended circuit with a tapped, link-coupled antenna coil. Although the popular pi-network can be used, the author prefers the link coupling system for mobile work because the final cannot be loaded unless the antenna is resonant. This is not necessarily true with a pi-network as evidenced by the hams who unknowingly load a length of coax line rather than an

antenna. The original final tuning capacitor is left ganged to the oscillator merely because removing it would wreck the entire dial drive assembly. Although an additional tank capacitor is used in the final, the original capacitor is connected in parallel with this to build a higher "C" tank and also to afford some degree of oscillator final tracking. If you want to be fancy, you can tailor the final tank coil and added tuning capacitor to achieve perfect tracking across the entire band. Since most of our operation is in the top 50 kc of the band, perfect tracking was not essential and frequency excursions of this order can be made without retuning the final (provided your loaded whip is resonant).

The modulator consists of a 12AU7 miniature dual triode with sections in parallel. (Yes, you can use 12SN7 or 12BH7 with no changes). This is driven by a resistance coupled 12AT7 speech amplifier. The carbon mike input circuit is the familiar grounded grid method which does away with the need for a mike current supply and mike transformer.

In the AM position, switch S1 performs two functions. First, the cathode of one of the final tubes is opened. This leaves us with a conventional, single-ended class C amplifier. Second, B plus is applied to the remaining tube's screen through the center-tap of the modulation transformer. End result: a screen modulated final! Throwing the switch to the other position restores the rig to DSB. A few minutes with the schematic will make this clear.

Construction

The original 1626 oscillator circuit is left intact and the output coupling link feeding the 1625 grids is reconnected to the buffer grid. The 6AK6 buffer, 12AT7 speech amplifier and 12AU7 are squeezed into the rear apron space formerly occupied by the crystal socket and indicator tube. The 0A2 voltage regulator sits just behind one of the 12DQ6's. No special precautions in construction are required other than the usual one of shielding long audio leads to prevent r-f and/or audio feedback. Power is supplied at the rear via a Jones plug while phono connectors are used for antenna connections. The original oscillator dial can be covered with paper and new calibrations inked in, or, it can be replaced with a disc of thin aluminum suitably marked.

Octal sockets are needed for the 12DQ6's and these are secured to a sheet of aluminum which covers the area formerly occupied by the 1625's. Removal of the final padder capacitor leaves room for the modulation transformer underneath. As can be seen from the photographs, the original tank coil and antenna roller coil assembly are removed to make room for the meter, antenna coupling switch and final tank tuning capacitor. Naturally, it is not necessary to follow this exact order of construction. Just make your own parts fit the available space. Note also that control circuits are not shown. Your pet ideas are probably better than mine so why complicate the schematic?

With the 250 volts applied (hold off with the 600 volts) and the oscillator perking at 3.9 mc, peak the buffer coil while measuring grid current at either of the final grid jacks. A check of both grid currents will probably show that one is higher than the other. Observe the higher current and adjust the mica grid trimmer to reduce this to a value approximately equal to the lower current. Repeak the buffer coil for maximum and again check both currents. If they are still not equal, repeat the above juggling process until they are. Once this is done, no further adjustment of drive is required to cover the phone band. With the values of grid resistors shown, the grid current should run between 1½ and 2½ ma per tube when the final is loaded. Somewhat better cross-over characteristics are obtained on a bow-tie scope pattern with the lower value grid current but this is not a really critical parameter. The amount of grid drive can be adjusted by selecting various 6AK6 screen dropping resistors. If the drive is adjusted for 2½ ma with the buffer coil peaked at 3.9 mc a drop of about ½ ma. will occur at the band ends. If the grid currents drop off

much more than this, the "Q" of the buffer coil is probably too high and this can be corrected by resistance loading with a one watt resistor of the order of 10K to 68K.

With grid drive adjusted and balanced, apply 600 volts to the final and throw the DSB-AM switch to AM. Resonate and load the final in the usual manner. In this position 80 ma is plenty and this represents 48 watts input to the single screen modulated final. Flipping the switch to DSB should drop the plate current to around 20 ma and speech should kick this up to around 150 ma or more. When the audio gain control is set to produce this amount of peak current on DSB you will find that changing over to the AM position will result in over-modulation. This can be corrected, naturally, by backing off audio gain, or, if you are as lazy as the average mobileer, you can just back away from the mike slightly.

Operation

DSB is used in the same manner as SSB, so by all means arrange things for push-to-talk so the voice controlled stations don't have to wait for you to fumble around with switches. Instant receiver recovery is necessary and you may need an extra set of relay contacts to short out the B plus that lingers in the filter capacitors of the exciter when the button is released.

In the QSO's we have had with SSB stations, none realized that both sidebands were being transmitted until we spilled the beans. Reports on speech quality were uniformly good and some glowing comments were received on the stability of the signal under mobile conditions. Signal strength reports were generally one or two "S" units higher than we used to get with the 60 watt AM rig. Perhaps the one frustrating feature of DSB or SSB mobile is the round-table with three, four or more stations involved. This requires a prodigious memory for calls and handles, plus log-keeping ability, while still maintaining the car on an even keel!

5.6 A 6146 DSB BALANCED MODULATOR

Fig. 5.6-A shows a balanced modulator that is used by K2GZT, with the major components described. Although the circuit is somewhat incomplete in-so-far-as a construction project is concerned, it does serve to illustrate how the popular parallel 6146 transmitter (B & W, Heath, Johnson, etc.) could be converted to a DSB rig. The only real modifications are in the grid circuit, where a push-pull configuration is required. The 12BH7 serves as a modulator and is supplied with push-pull audio, about 20 volts peak-to-peak.

The operating conditions are listed in fig. 5.6-B. Note that the plate voltage specified (1200 v.) is much higher than the usual maximum rating. It is in line with the usual practice in a DSB balanced modulator, of operating the tubes at a dc plate voltage twice the recommended value for plate modulated class C amplifier service. Thus the 1200 volt value is equal to the positive peak modulating voltage. If you are converting an AM transmitter to the circuit shown in fig. 5.6-A, use what-ever plate voltage is available, up to 1200 volts.

R-F Drive

Approximately 4 to 5 watts of driving power should be available for the balanced modulator even though the 6146's actually do not require all of this power for proper operation. The exciter output must be free of spurious output frequencies, and have excellent frequency stability. Preferably, any variable frequency oscillator used with the DSB modulator should not have more than one kc of warm-up drift during the first few minutes of operation, and should be capable of staying within 50 cycles of the desired operating frequency, once it has stabilized.

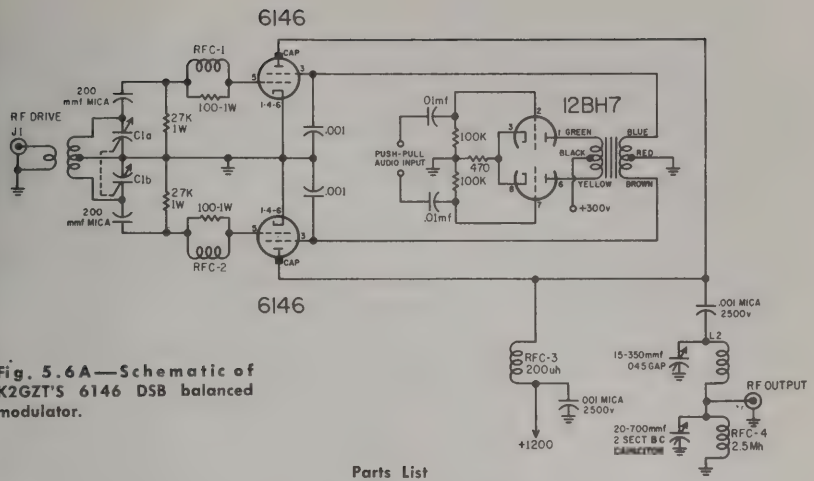


Fig. 5.6A—Schematic of K2GZT'S 6146 DSB balanced modulator.

Parts List

C1a, C1b — 10 — 100 mmfd. per section, two section variable capacitor
J1, J2—Chassis type coaxial connectors
L1—3.5 mc. B & W JVL-80, 7.0 mc. B & W JVL-40, 14.0 mc. B & W JVL-15, 21.0 mc, B & W JVL-15.
L2—3.5 mc. 6.5 uh, 18 turns #16 wire, spacewound 8 turns per inch, 2¼" long, 1½" in diameter.
7.0 mc. 3.2 uh., 13 turns, #16 wire, spacewound 6 turns per inch, 2-1/16" long, 1½" in diameter
14.0 mc. 1.6 uh., 9 turns, #14 wire, spacewound 4 turns per inch, 2¼" long, 1½" in diameter.
21.0 mc. 1.08 uh., 7 turns, #14 wire, spacewound 4 turns per inch, 1¾" long, 1½" in diameter
RFC1, RFC2—10 turns #22 enam. closewound and self-supporting, ¼" in diameter
RFC3 — 200 uh., r.f. choke (Rapar RL-111)
RFC4 — 2.5 mh. r.f. choke (Miller #6302)
T1 — Driver transformer, 5.2:1 turns ratio, reverse connected (Thordarson 20D79)

Mechanical Layout

As stated earlier, no particular construction information is provided. If the 6146 balanced modulator is a conversion of an existing AM transmitter, the layout will be dictated by the design. The only problem occurs in fitting in the push-pull grid tank circuit: This push-pull circuit should be link coupled to the exciter tube tank coil to avoid capacitive unbalance. If the 6146 final is to be constructed "from scratch" the usual practice of short leads in the r-f circuitry should be followed. Try to keep the layout of each 6146 grid circuit identical so that stray capacities will be the same, thus maintaining balance. The final amplifier layout for any of the popular 100 watt class transmitters (DX-100, Viking, etc.) may be used as a guide for this circuit.

DC Plate Voltage	1200 volts
DC Screen Voltage	0 volts
DC Control Grid Bias (other than signal bias)	0 volts
DC Plate Current (no modulation)	25 milliamperes
DC Plate Current (modulation)	100 milliamperes
Meter Peak Power Input	240 watts

Fig. 5.6B—Chart of operating conditions for the 6146 tubes in a DSB balanced modulator circuit. The current figures shown are per tube and are consistent with good linearity.

5.7 SYNCHRONOUS DETECTION

Synchronous detection is similar to the phasing technique used for SSB reception *except* that the locally generated carrier is *phase locked* with information derived from the sidebands of the transmitted signal. Note that the carrier, if transmitted, is ignored by the receiver. Synchronous detection takes advantage of the inherent diversity feature of all double sideband emissions, since the detector combines the intelligence in both sidebands. This is of value under conditions of selective fading.

W2CRR, John Costas, has shown that maximum receiver performance is obtained by synchronous or coherent detectors. With detectors of this type, the audio output of both sidebands add to reinforce one another when the local oscillator is maintained in proper phase relationship. When the local oscillator phase is 90° from the proper phase, the audio output contributions of the upper and lower sidebands cancel each other and an audio null results. We may use this sideband relationship to reinsert a demodulating carrier at the correct frequency and phase to extract the intelligence from the sidebands.

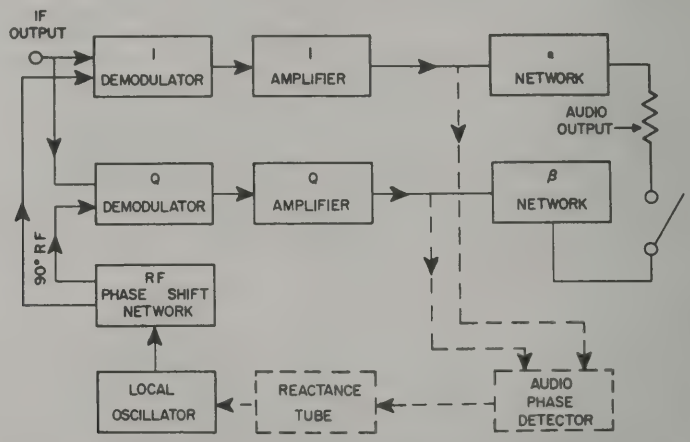


Fig. 5.7A—Block diagram of a Synchronous Receiving Adapter described in the text.

Basic Elements

The elements of a synchronous detector are shown in fig. 5.7-A. The signal input is applied to two basic demodulators and the local oscillator voltages are applied to each demodulator in *phase quadrature* (90° apart). Assume momentarily that the "I" channel local oscillator injection voltage is the same phase as the carrier (transmitted or suppressed) component of the AM signal. Then, the in-phase or "I" channel will contain the demodulated signal while the quadrature or "Q" channel will contain no audio as its injected local oscillator is shifted 90°.

Now, if the local oscillator drifts slightly, the "I" channel will be relatively unaffected but the "Q" channel will produce some audio. This will have the same polarity as the "I" channel audio for one direction of local oscillator drift, and opposite polarity for opposite direction of local oscillator drift. The "Q" channel level will be proportional to the oscillator drift for small errors. By simply combining the "I" and "Q" audio in an audio phase discriminator, a dc control signal is obtained. This control voltage tunes the oscillator via the reactance tube and returns or "locks" the oscillator to the correct phase, where audio is present only in the "I" channel.

The audio phase detector delivers a dc voltage only when the "I" and "Q" signals have in-phase components. Since "I" and "Q" audio will be in phase quadrature in any case, where the like sideband components do not exist on each side of the carrier, the phase detector provides no AFC voltage for SSB or CW signals. The phase detector is therefore unaffected by heterodyne interference. With no phase lock, audio appears in both "I" and "Q" channels and the 90° audio phase shift networks will permit SSB reception by the phasing method.

If the "I" and "Q" outputs are taken through alpha and beta networks respectively, interference rejection may also be obtained. When locked on a signal containing interference in the lower sideband, for example, "I" channel produces audio resulting from both sidebands and lower sideband interference, while "Q" channel contains only interfering audio. Phase cancellation by combining the two audio outputs will remove the interference while still adding the desired information in both sidebands. Other combinations are obviously possible.

A few not too demanding precautions must be observed in order to obtain optimum performance from the synchronous detection process. They are:

1. Intermodulation distortion in the demodulators must be low.
2. The "I" and "Q" channels must have identical phase shift characteristics. Use of matched resistors and capacitors is best, but 5% tolerance components are adequate.
3. The "I" and "Q" channel local oscillator injection voltages must be in quadrature.
4. Audio operated AGC must be used in order to prevent overdriving the audio phase detector.
5. The frequency response of the "I" and "Q" inputs to the phase detector should drop at least 12 db per octave below 300 cycles.

Although local oscillator stability must be reasonably good, stability sufficient for SSB reception is more than adequate for synchronous detection.

The block diagram of such a device is shown in fig. 5.7-A. The schematic of fig. 5.7-B contains all the elements of the block diagram plus the addition of an AGC circuit. The adapter may be connected to the receiver via the second detector tube socket on many receivers. The i-f is brought out; audio and AGC voltage (derived from audio, since carrier is ignored) are fed back to the receiver, and the "S" meter will indicate sideband strength.

Product detectors, V1 and V2, were chosen for the "I" and "Q" demodulators in lieu of exalted carrier diode detectors in the interest of linearity. The "I" and "Q" audio amplifiers, V3a and V3b, are conventional. The audio phase detector is of greatest interest and is the heart of the system. It is fed with audio from the "I" and "Q" channels through the phase inverters V4a and V4b. With no audio present in the "Q" channel, no dc output appears at TP3. If audio appears in the "Q" channel, indicating incorrect tuning of the local oscillator, dc of the correct polarity to retune the oscillator appears at TP3 with the diodes connected as shown. The local oscillator and reactance tube circuitry are conventional. T1 was made from a 455 kc. i-f transformer. One winding was replaced with 100 turns of #36 enamel wire on a ½ inch diameter sleeve and the coupling adjusted to give -6 to -8 volts across R38. The reactance tube sensitivity should be 1.0 kilocycle per volt. If another oscillator coil is used it will be necessary to adjust the reactance tube sensitivity. Reactance tube sensitivity will be approximately proportional to the reactance of C30. If the oscillator should fail to oscillate with the specified coil, it may be necessary to increase the number of turns on the feedback winding. The 90° local oscillator phase shift is obtained from two 45° phase shifts through C24, R35, R36, C22 and C23. Oscillator injection voltage to each demodulator grid should be about 1.0 volt. For operation at an i-f other than 455 kc the capacity values should be changed in an inverse proportion to frequency. Tubes V7 and V8

and the associated audio networks are not essential to the synchronous detection process but facilitate alignment and add to the usefulness of the adapter. The networks are conventional 90° audio phase shift networks familiar to SSB enthusiasts and are available commercially or may be assembled from measured components. Assuming the adapter is correctly aligned, the networks provide "Q" channel interference audio 180° out of phase with either the upper or lower sideband of "I" channel audio, depending on the position of switch S2. When the local oscillator is synchronized or "locked" on a signal, audio resulting from both upper and lower sidebands appears at "I" channel output and no sideband audio appears at "Q" channel output. If interference appears in the lower sideband it will result in interfering audio in both "I" and "Q" channel lower sideband outputs. By switching S2 to REJECT LOWER, all interfering signals in the lower sideband will cancel in the combined "I" and "Q" outputs and all desired information present in the lower sideband will be retained for addition with upper sideband audio.

Upper sideband interference may be similarly rejected by switching to REJECT UPPER. Remember, "Q" channel has no components of the desired double sideband signal, so it cannot cause cancellation of the desired audio from either sideband. If the local oscillator is "unlocked" by grounding the audio phase detector output (AFC OFF), audio appears in both "I" and "Q" channels and the adapter becomes an SSB converter. SSB, or either sideband of a DSB, AM, NBFM, or PM signal may be selected. This interference rejection feature may also be used to improve cw reception. The audio operated AGC circuit, V6, works well with all types of signals. The attack time is controlled by R50 and C34 and the release time is controlled by C34 and the resistance looking back into the receiver. The time constants may be modified slightly by the receiver constants. AGC as used here was found superior to conventional AVC, as AVC regulates on carrier information and this circuit operates on audio output independent of modulation percentage.

Prior to connecting the adapter to a receiver, the BFO should be adjusted to the center of the i-f passband by tuning it so that the lowest pitch noise is heard. Do not readjust. Remove the second detector diode from the receiver and connect the adapter input to the socket pin which is connected to the last i-f transformer secondary. This connection takes i-f from the receiver and feeds AGC voltage back in. If the receiver AVC voltage is developed somewhere other than in the envelope detector, the AGC voltage should be applied by direct connection between C34 and the receiver AVC line. The adapter audio output should be connected to the receiver phono input or audio amplifier.

Adjust the tuning of T1 to produce zero beat with the receiver BFO. This aligns the adapter to the center of the receiver i-f passband to assure equal amplification of both sidebands. The receiver BFO may now be detuned approximately one kc and, with AFC OFF, the INTERFERENCE REJECTION should be set to the position giving the least audio output. Adjust C22 and R67 in succession to null the output. If the alpha and beta circuits were not included, a scope could be used to adjust for 90° phase difference between the "I" and "Q" audio signals (or local oscillator injection). Up to this point the alignment may be compared to that for a "signal slicer". If the interference rejection and SSB reception features are not desired, V7 and V8 should be omitted and audio output taken from R10.

To align the phase detector, the BFO should be OFF and the receiver tuned to an unmodulated carrier so that a one kc beat note is heard. Adjust the receiver r-f gain to produce about four volts of audio on the plate of the audio amplifier, V3b. Short R11 to ground, adjust R15 for dc null at TP1, adjust R16 for dc null at TP2, and adjust R31 for minimum ac at TP3. Turn the receiver AVC ON and advance the r-f GAIN control. The S meter should move up-scale, but there should be little change in audio output level. The AGC delay voltage is on Pin 8 of V6 and should be about plus 20 v dc. The adapter should now be ready for reception.

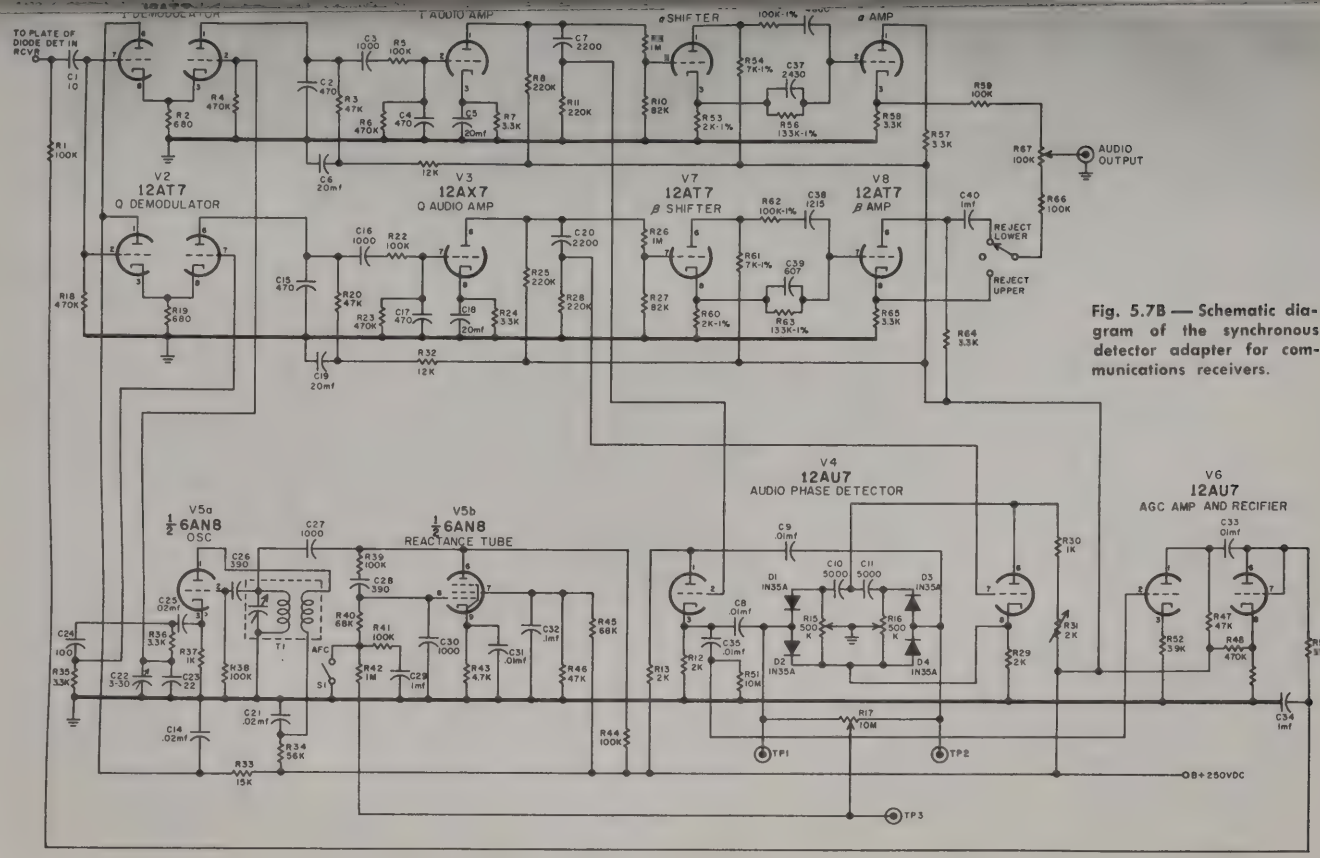


Fig. 5.7B — Schematic diagram of the synchronous detector adapter for communications receivers.

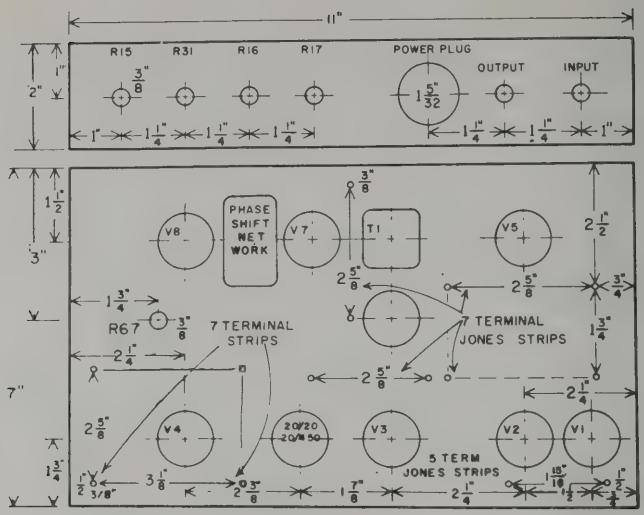


Fig. 5.7C—Chassis layout for the Synchronous Adapter.

In general, the front panel controls of the receiver will perform the same functions as before the conversion. With the receiver AVC ON, and the adapter AFC ON and INTERFERENCE REJECTION at ZERO, an AM station is received by tuning to zero beat. With the constants specified the oscillator will "lock" if the receiver is tuned to within 100 cycles of the center of a DSB signal and undistorted audio will result. If heterodynes cause the oscillator to shift frequency or output is garbled, the phase detector connections and the adjustments should be rechecked. Tuning accuracy will be greater than required on conventional AM receivers but the degree of tuning accuracy necessary for SSB reception will not be required (except when SSB signals are being received). For this small sacrifice in tuning tolerance you will gain: (1) vastly improved phone reception under conditions of selective fade; (2) ability to employ the simple and efficient DSB transmitter (even at VHF); (3) interference rejection; (4) ability to receive DSB, AM, NBFM, PM, SSB and CW without gadgets; (5) better performance due to inherent diversity of two sidebands properly received; and (6) eventual heterodyne free ham bands after all have converted to simplified and carrier free transmitters.

5.8 A 100 WATT DSB ALL BAND TRANSMITTER

The author would like to express his thanks to Globe Electronics, and particularly to Mr. Alan McMillian, for permission to reproduce the schematic and technical data on the Globe Sidebander, Model DSB-100. The DSB-100 is the first commercially manufactured double sideband rig for Amateur use and is worthy of special mention. It may be used directly into an antenna system, "barefooted" that is, or may be used to drive a high power linear amplifier. A variable r-f output control and speech clipper assures the operator that the linear amplifier will not be overdriven even on the highest voice peaks. Fifty db of carrier suppression may be obtained by careful adjustment of the internal balancing trimmer. A pi-network will allow a great variety of antennas to be used with maximum efficiency.

Fig. 5.8A—Schematic diagram of the Globe "Sidebander" Model DSB-100. The transmitter runs 100 watts peak power input, and contains a very effective speech clipper.

Theory of Operation

A 6CL6 tube operating as a modified Colpitts crystal oscillator supplies r-f excitation for the transmitter. Eighty and forty meter crystals will provide output on all Amateur bands 80 through 10 meters. The output of the oscillator is developed across broadband plate coils L1 and L2, resonant at 40 and 80 meters respectively. Cathode keying of this stage is provided for VOX or CW operation.

A 6CL6 is also employed as a buffer or frequency multiplier. Output from this stage is on the desired operating frequency and appears across the push-pull grid circuit L3- C10. A balancing trimmer C11 is set so that its capacity is equal to the tube and stray capacity on the other end of the tank coil. The plate coils are progressively shorted from the center towards each end to obtain the proper amount of inductance for each band.

The final amplifier consists of a pair of high perveance 6DQ6's connected in the typical DSB balanced modulator configuration. The grids are connected in push-pull, the screens are modulated in push-pull, and the plates are connected in parallel. For AM or CW operation, one of the 6DQ6's is disconnected and B plus is applied to the screen of the other tube, in series with the modulation transformer. The cathodes of the final amplifier tubes are also connected to the key jack. CW operation merely entails plugging a key in the jack, in the usual manner. The VOX relay is also used to open this circuit during periods of standby. Metering of the final amplifier grid and cathode circuits allows the operator to tune properly and to monitor these currents to assure normal operation.

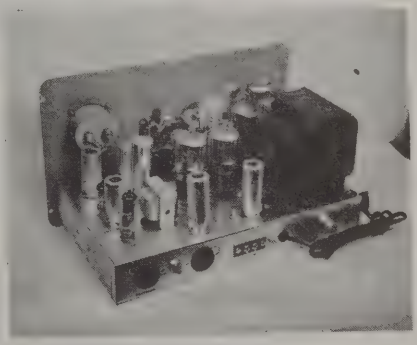
A dual purpose tube, type 12AX7, serves as the microphone amplifier, second speech amplifier and the tone oscillator. The first half of the 12AX7 (V5a) functions as a convention r-c coupled audio amplifier when the TONE switch is in the OFF position. When the tone switch is placed in the ON position, the grid of V5a is connected to the plate through the phase shifting network, and the stage oscillates. The output is an excellent sine wave at a frequency of approximately 600 cycles. The audio output (either tone or voice) is capacity coupled to V5b, where it is further amplified to a sufficient level for audio clipping. Audio from V5a is also coupled to the grid of V6a, the VOX amplifier. Output from this stage drives an accessory voice relay.

Globe Electronics enters the sideband field with their Model DSB-100. This DSB/AM/CW rig runs 100 watts peak input with a pair of high perveance 6DQ6's in the balanced modulator. A very effective speech clipper greatly increases talk power.

The audio from V5b is coupled to the speech clipper, which functions in the following manner: The plates of V7 are connected to a positive bias voltage which is taken from the high voltage line. The bias is adjusted so that when positive audio peaks exceed the fixed bias voltage at one diode, it stops conducting. Negative audio peaks are similarly affected by the other diode. Adjustment of the bias voltage or the strength of the incoming audio signal governs the degree of clipping. Normal current flows from cathode to plate and there is no clipping when the bias voltage is higher than signal voltage as a conduction path exists in the first diode of V7. When the positive peak audio signal equals or exceeds the bias voltage, the tube stops conducting and clips that amount of the audio wave that exceeds the bias voltage. Negative audio peaks work similarly on the second diode section which is biased in reverse to the first diode. Clipping normally produces square waves, which in turn produce odd harmonics. An LC filter follows the diodes to attenuate these odd harmonics.

The normal action of a clipper stage automatically attenuates the audio output to a degree governed by the setting of the clipper, therefore, an audio amplifier is necessary to amplify the audio to its original level. This is the function of V6b, which in turn, drives the modulator tube V8. The modulator is a transformer coupled pentode amplifier which serves to drive the final amplifier screen in push-pull.

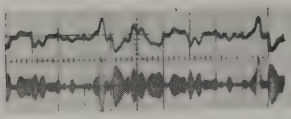
The power supply consists of a transformer operated full wave rectifier utilizing a choke input filter system. The high voltage for the r-f exciter is regulated to within 15% from full to no load through the use of variable resistance "Globar" type units (R8- R9). An accessory socket on the rear apron provides 6.3 vac, 600 vdc, 300 vdc, key, audio output and ground.



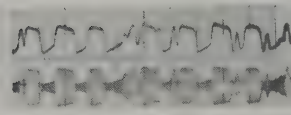
Inside view of the Globe Electronics DSB-100. The audio/speech clipper section is located on the left side of the chassis. The two shielded tubes along the rear apron are the 6CL6 oscillator and frequency multiplier. The final amplifier and band change network is located near the center of the chassis. A husky power supply completes the unit.

Tune-up and Operation

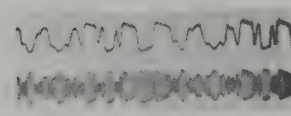
As with other DSB rigs, the DSB-100 is very simple to tune up and operate. Once the mike, key (or VOX relay), and antenna are connected, and the power applied, place the bandswitch on the desired band and switch the meter to the grid position. Switch to the DSB mode, and tune the grid drive for maximum grid current. Now, switch the meter to plate position and turn the tone generator on. Increase the audio volume until the plate current rises to 130 ma. (It will read about 30 ma when "idling"). Tune the final plate for minimum current. Then, advance the Gain control until the pointer is at 2 o'clock. Adjust the plate tune, and plate load controls alternately until the current in the "bottom of the dip" is 130 ma. The transmitter is now set up for normal DSB operation. The plate current should idle at 30 ma and kick up to 130 ma with modulation. The reserve gain in the audio amplifier will allow full input even with a high degree of speech clipping. The clipper circuit in the DSB-100 is very effective as evidenced by many Q5



(a) Voice signal
(b) Resulting DSB RF waveform
} no clipping



(c) Voice signal
(d) Resulting DSB RF waveform
} with clipping



(e) Voice signal
(f) Resulting AM RF waveform
} with clipping

Speech waveforms in a DSB/AM transmitter with and without speech clipping. Note the higher average power in waveform d and f, as compared to that in b.

contacts under adverse conditions. When constructing DSB rigs, or converting existing transmitters, it would be advisable to include a speech clipper similar to the one used in the DSB-100. The very effective increase in "talk power" will more than compensate you for the extra time necessary to construct the unit.

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6

Linear Amplifiers

IN THE foregoing chapters we have shown how it is possible to generate single sideband signals by either the filter or phasing method, or DSB signals with simple balanced modulators. The remaining problem is to amplify these signals to a high enough level so that they can hold their own in the presence of ham band QRM.

Let us examine the quality of the SSB signal that has been generated in these exciters. Any of the units, when properly aligned and operating conservatively into a proper load, is capable of 40 db attenuation of the undesired sideband. This is the sideband that has been so painstakingly filtered or phased out. There are also other sorts of signals that appear, not only in the spectrum occupied by the undesired sideband but in the region of the transmitted sideband as well. These signals are the products of distortion in heterodyning and amplification of our sideband signal. These distortion products are not necessary and are just so much garbage as far as conveying information is concerned. Most SSB exciters will have distortion products that are approximately 50 to 60 db below the peak value of the transmitted sideband voltage. This is considered to be a very clean signal. If we could just retain these attenuation ratios, everything would be just dandy. Unfortunately, we can't, and upon trying to do so hangs the tale that follows.

6.1 LINEAR AMPLIFICATION

Let us look into the more important aspects of linear amplification. First, what is a linear amplifier? It is one that faithfully reproduces all of the input signal amplitude variations in the amplifier output circuit. What classes of amplifiers are linear? What are some of the examples? Class A amplifiers are the best known of the lot. The average plate current remains constant, the grid voltage never (no, never) swings into the positive region, and the distortion products are so low as to be negligible. Examples? The r-f and i-f amplifiers in your communications receiver are operating class A. So are the microphone preamp and low level stages in your modulator. The efficiency is quite low—the peak efficiency being in the order of 25% to 35%. Most cathode biased linear output amplifiers in sideband exciters are also operating class A.

Class AB_1 amplifiers are somewhat like class A amplifiers in that the grid never swings positive and therefore never draws grid current. The efficiency is higher, the maximum is approximately 55%, but the distortion products are worse—not much, but enough to be noticeable. One might suspect that if the two classes of amplifiers discussed above do not draw grid current there is no grid driving power required. Actually, this is not true because we must supply the grid circuit and coupling circuit losses. Fortunately, this does not amount to much—a fraction of a watt, generally. The driver must furnish enough grid voltage swing to operate the stage.

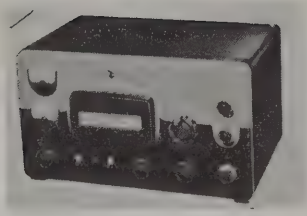
The class AB_2 amplifier is a cross-breed of the AB_1 and the full class B stage. The average plate current (as indicated by the plate current meter) will kick upward and the grid will swing into the positive voltage region for a portion of the excitation cycle and, therefore, draw grid current. Generally the no-signal plate current is higher than encountered in full class B stages.

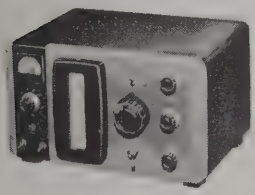
The bias on the control grid is usually set so that the idling plate dissipation of the tube is approximately half of the maximum specified for the tube. When signal is applied, the input goes up, and there is power delivered to the output circuit. The remaining power not lost in the output tank circuit or coupling circuit appears as the operating tube plate dissipation. This operating dissipation does not exceed the maximum rating—at least not for long. More on this later.

From the foregoing discussion, I believe that you can sense that the class B amplifiers are the ones that get "horsed around" a bit. The plate current is run at a lower idling value by increasing the bias and the grid is usually driven farther into the positive voltage region. This causes it to draw more grid current and therefore, requires more grid driving power. The plate current swings over a greater range than do the AB_1 or AB_2 amplifiers. Correspondingly, the efficiency is higher with a theoretical maximum placed at about 78%. However, the possibility of distortion products is greater.

The Technical Material Corp. Model PAL-350 linear amplifier uses two 4X250B tubes operating in class AB_1. This unit will deliver 350 watts of PEP power between 2 and 32 mc with only 0.1 watts of driving power.

The Johnson Viking "Thunderbolt" runs 2000 watts PEP power input and requires only 20 watts of drive. Two 4-400 are used in class AB_2. The "Thunderbolt" has a terminated input switch position similar to the one shown in Fig. 6.4-E, which eliminates the need for grid tuning adjustments.





The Johnson Viking "Courier" uses two 811 tubes operating class B. The unit is rated at 500 watts PEP power input, with 5 to 35 watts driving power required. A pi-network tunes 3.5- 30 mc and will load antennas between 40 and 600 ohms.

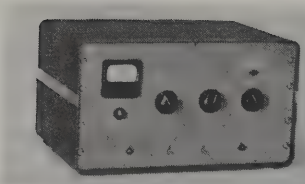
6.2 SPLATTER? WHO ME?

The term "distortion products" keeps popping up all the time. Just what are they? The popular terms are splatter, monkey chatter, crud, and a few other names that cannot be printed here. I am sure you are all familiar with the terms, and the effects of splatter, but you may not understand how it can be generated in a linear (so-called, that is) amplifier.

A true linear amplifier will produce output signals that are amplified replicas of the input signal. If the grid signal varies in amplitude between the limits of 1 to 2 volts, the output of a linear amplifier will vary between the limits of say 100 to 200 volts. As you can see the stage has a voltage gain of 100. But suppose that the output signal does not quite make the grade and varies between 100 and 185 volts with a 1 to 2 volt grid swing. That, gentle reader, is non-linearity. Proceeding with this line of reasoning, assume that the linear uses one tube in conjunction with an output circuit parallel tuned to the desired operating frequency. The non-linear condition will cause a fundamental frequency plate current pulse to flow in the tank circuit. In addition, harmonic signal currents (2nd, 3rd, etc.) will also appear in the tank circuit. If the tank circuit has a good operating "Q", the harmonic signals will not be transferred to the output circuit. However, if the input signal is made up of more than one frequency (such as a group of frequencies representing the human voice), a peculiar thing takes place. Each of these frequencies, when passing through the non-linear amplifier at a signal frequency of say 4 mc., will have harmonic frequencies near 8 mc., 12 mc., 16 mc., and so on. Such a stage is said to have *intermodulation distortion* and distortion products are the net result.

At this point specific numbers would be very useful to examine the distortion products more closely. Assume a suppressed carrier frequency of 4000 kc, and the generation of the lower sideband. Pick two audio input frequencies for convience—say 1000 cycles and 2000 cycles, since they are nice "round" numbers. The two modulation frequencies will net us two sideband signals at 3999 kc and 3998 kc respectively. Now, let's apply these two signals to the non-linear "linear" and see what takes place.

The fundamental signals, 3999 kc and 3998 kc will certainly appear in the output circuit. For all practical purposes the harmonic frequencies will not appear because the tank circuit has a good "Q" and can discriminate against these unwanted har-



The Gonset GSB-101 linear amplifier is rated at 750 watts peak power input and is designed to work with exciters supplying 75-100 watts of drive, such as the GSB-100.

monics. Never-the-less, the harmonics will make their presence known, and in a very unusual manner. A scratch pad will reveal that the 2nd harmonic of 3999 is 7998 kc and of 3998 is 7996 kc. Since our amplifier is non-linear it is capable not only of amplification, but also of heterodyning. In Chapter II, it was stated that any non-linear device could be used as a mixer. It was further stated that for good mixing, the tube must be operated on the non-linear portion of the E_g-I_p curve.

For this explanation, our only concern is for the mixtures of the fundamental frequencies and *any* of the harmonic signals, or mixtures between the harmonic signals themselves, which may fall near 4.0 mc. All other combinations will be disposed of by the selective properties of the tuned output circuit.

CASE 1—Mixing the 2nd harmonic of 3999 kc (7998 kc) and the fundamental signal 3998 kc. The sum mixture will be 11,996 kc and will not appear in the output. The difference frequency is 4000 kc and will appear in the output.

CASE 2—Mixing the 2nd harmonic of 3998 kc (7996 kc) and the fundamental signal frequency of 3999 kc. The sum mixture will be 11,995 kc and will not appear in the output. The difference frequency is 3997 kc and will appear in the output.

CASE 3—Mixing the 3rd harmonic of 3998 kc (11,994 kc) and the 2nd harmonic of 3999 kc (7998 kc). The sum mixture will be 19,992 kc and will not appear in the output but the difference mixture is 3996 kc and this will appear in the output.

CASE 4—Mixing the 3rd harmonic of 3999 kc (11,997 kc) and the 2nd harmonic of 3998 kc (7996 kc). The sum mixture again will be rejected but the difference mixture of 4001 kc will appear in the output.

Only the 2nd and 3rd harmonics of the two signals fed into the amplifier have been considered. The 4th, 5th, and so on might be significant also, but the explanation shown illustrates the point very well. The amplifier is fed only two frequencies and yet, look what comes out the antenna terminals! The original signals, 3999 and 3998 kc are present, of course. But, the following *new* signals are also there: 4000 kc (this happens to be the suppressed carrier frequency), 3997 kc, 3996 kc, and 4001 kc. All but the last signal are at or below the carrier frequency, but the 4001 kc product is in the upper sideband. This is in the region that has been carefully suppressed in the exciter. In effect, the non-linear characteristic of the amplifier has caused the un-wanted sideband to be re-inserted, in addition to the splatter in the lower sideband region. It should also be pointed out that sideband suppression is directly proportional to the linearity of the amplifier. Therefore it is reasonable to assume that the amplifier linearity can be adjusted by observing the sideband suppression as described in Chapter IV.

Amplifier non-linearity can cause spurious signals above and below the suppressed carrier frequency. When listening to such a signal, the modulation will appear to sound harsh. A quick check of the other sideband will usually disclose the presence of distortion products. If the signal "stinks to high heaven", you will be able to detect splatter for many kc in either direction from the suppressed carrier frequency. The VK/ZL boys have a quaint way of expressing it, when your signal is not all that it should be. Better check things, if they inform you that your signal has "whiskers" on it.

6.3 DISTORTION — CAUSES AND CURES

The picture has been painted rather black so far. The cause and effect of distortion products has been shown, but what about the cure? The following sections deal with the different localized causes of distortion and how to overcome them.

Improper Grid Bias

If the grid bias is too high, it will cause the tube to operate on the non-linear "knee" portion of its characteristic curve, at the cut-off end. If the bias is too low, it will cause the idling plate dissipation to be excessive. A visual check will usually indicate such an unhealthy condition. There is a handy rule of thumb that you can use to set the bias on a linear amplifier. It should be adjusted so that the no-signal plate dissipation is at least ⅓ to ½ of the maximum rating. Another grid bias requirement is that the bias voltage must remain constant under all operating conditions. This dictates that there be no resistance in the grid circuit whenever any grid current is drawn. The bias supply, whether electronic or battery, must have a minimum of internal resistance (or impedance, as the case may be). There is one obvious way to dodge this problem and that is to use zero bias tubes. More about this later.

Improper Regulation of the Driving Signal

If the amplifier being considered never draws any grid current (as in class A or class AB_1) this is of no concern. However, if the grid, at some time during the excitation cycle swings positive, it will draw grid current. The load on the driver will increase sharply and cause the grid signal voltage to drop from its otherwise no-load value. This trouble is not really in the grid circuit but is dependent on the so called "internal resistance" of the driver stage.

It is possible to maintain the grid voltage during periods of heavier loading (during grid current periods) by lowering the internal resistance of the driver stage. This can be accomplished by lowering the tuned circuit "Q" (lower L to C ratio) or by "swamping" the driver stage plate circuit with a resistor, but this of course lowers the driver output. The best practice is to generate about four times the power necessary to drive the final and swamp the remaining three quarters of the power with a resistor across the driver tank. Since the driver stage is usually low power, this procedure is not too wasteful.

Improper Grid Drive

The grid can be over driven so heavily that the linear will "saturate". This over driving will cause flattening of the peaks of the output waveform which produces distortion products in large quantities. Conversely, the grid can be getting too little drive and the exciter may be trying to deliver more power than its capabilities. Thus the distortion may be generated in the exciter and not in the final.

Plate Circuit Distortion

The plate tank is the device that transfers the signal energy from the final tube (s) to the antenna circuit. The operating "Q" of the plate tank must be high enough so that the harmonics are attenuated, but not so high that the plate circuit efficiency falls off. The generally accepted limits of the loaded "Q" are from 12 to 15.

Distortion can also be created in a linear amplifier if the loading is incorrect. If the stage is loaded too lightly, the amplifier will be driven into saturation much sooner than normal and the output power will be reduced considerably. If the loading is too heavy, the stage will not saturate easily, but the output power will be lower than that obtained at optimum coupling. Use some sort of output indicator when tuning up a linear amplifier, such as an r-f ammeter or field intensity meter. Adjust the coupling for maximum output at some high value of power input. The plate tank circuit is discussed in more detail under "Design Considerations".

The P & H Electronics Model LA-400B uses four type 1625 modified tubes in a grounded grid configuration to supply 400 watts PEP input with 20 watts drive. The tubes (12 volt 807's) are modified so that the beam forming plates may be grounded, thereby increasing the amplifier stability.

6.4 TUBES AND CIRCUITS

There are several schools of thought on the tube matter. There are the two basic divisions, tetrodes versus triodes. Then in the triode class there are again two main groups of adherents: (1) those who prefer the zero bias tubes with their freedom from biasing troubles and severe swamping requirements, and (2) those who prefer to use the low-mu triodes with their high bias requirements so that high power may be obtained without going into the grid current region at all.

Each has its advantages and disadvantages. The zero-bias tubes will give higher stage efficiency (about 70%). There is no sharp transition from no-grid current to grid current region since it draws grid current as soon as even a small amount of grid excitation is applied. It will load the driver heavier at higher levels, but the sharp transient is missing and less swamping is needed than where a heavily biased tube is driven into the grid current region. The zero-bias stage is reasonably free of intermodulation distortion when properly operated.

The low-mu tube operating in class AB_1 requires practically no driving power, but will require a fairly high grid voltage swing in order to drive the tube to its full capabilities. The distortion products from this type of operation are very low—almost as low as class A operation. The only possibilities for trouble would be in the tube characteristic itself or in improper loading or drive as explained earlier. The stage efficiency is low (50 to 55%) but that is the price you pay for a cleaner signal. It is my own opinion that a clean signal is to be desired even at the expense of total output power.

There is "middle ground" in this matter, such as the tube that is driven into the grid current region for just a portion of the grid excitation cycle. This is the baby that you have to swamp heavily to give good grid signal-voltage regulation. This "middle ground" case unfortunately is not a compromise from the standpoint of distortion. Generally, it has more severe distortion present in the output than either of the previously mentioned triode cases.

This brings us to the next major class of tubes—tetrodes and pentodes. These, too, have their advantages and disadvantages. Their chief advantage is in the low driving power requirements and relatively low grid signal voltage required for full output. One of the chief objections to using the larger tetrodes is the necessity of a stiff regulated screen power supply. Ordinary voltage regulator tubes (VR-150, VR-105, etc.) will be suitable for many tubes while others will require the full treatment. This means that an electronically regulated supply using 6L6's, 6Y6's, or 6AS7's must be constructed, or a shunt regulated arrangement using a small transmitting tube and dropping resistor from the plate supply as suggested by W2AZW. The shape of the tube characteristic curves will probably have more effect on the distortion products in the output than anything else—if the circuit voltages are according to Hoyle. There are some tubes that just aren't suitable for use as linear amplifiers.

What Circuits? Here again everyone has their own preference. We shall try

C1, C2, C3—0.01 μfd., 600v., ceramic.
C4—120 μμfd., single section variable, air.
C5—0.002 μfd., 2000v., mica.
RFC1—50 turns #16 enam. on 1" form.
V1—Any zero-bias tube.

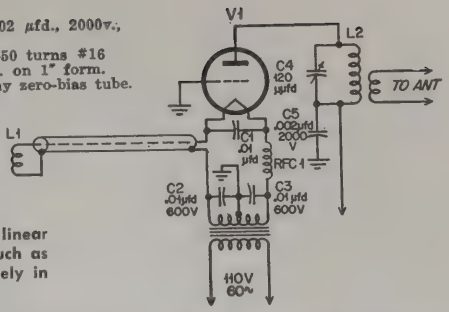
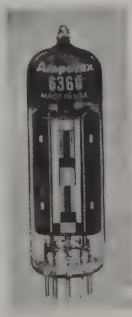


Fig. 6.4A—Grounded grid linear amplifier. Zero bias tubes such as the 805, 811, etc., work nicely in the configuration.

to review briefly what the various possibilities are. The two obvious general classes of circuits used are: (1) grounded grid, and (2) grounded cathode amplifiers. The grounded-grid amplifiers have some very attractive features. There is generally no tuned circuit needed for the input of the stage. There is no neutralization needed in the case of triodes, therefore the plate tank circuit can be single-ended, that is, no split-stator capacitor and center-tapped coil. If zero-bias tubes are used this further simplifies the problem. See fig. 6.4-A for an example of what can be done—note the minimum of parts needed. So much for the advantages, and now for the disadvantages.

Grounded-grid operation will require considerably more drive than use of the same tube in the conventional grounded cathode arrangement. This extra drive is not lost, however. It appears in the output circuit of the final amplifier as useful output. It is possible to get an output power from a grounded-grid stage that is greater than the dc input to the plate circuit of the amplifier. This would make the apparent efficiency greater than 100%. As mentioned this is because the driving power appears as output power. Don't get any ideas now—the FCC has taken care of what you are thinking about, with the amateur regulations regarding grounded grid operation. Still another point—never use a grounded grid stage as a driver stage where the driven stage reflects a changing load back into the grounded-grid stage because this changing load will in turn be reflected back one stage more into the driver's driver. Confusing, isn't it? Confusing or not, the results are bad—more distortion than is healthy.

Again referring to fig. 6.4-A, you will notice that a filament choke is necessary to keep the filament transformer capacity to ground from shunting the r-f driving voltage. The other alternative (to using a choke or tuned circuit in the filament wiring) is to procure a special low-capacity filament transformer that some of the surplus radar sets were blessed with.



Amperex makes an interesting line of tubes for single sideband service. The 5894, in particular, has been found to be exceptionally linear and is becoming quite popular for commercial and "home-brew" rigs.

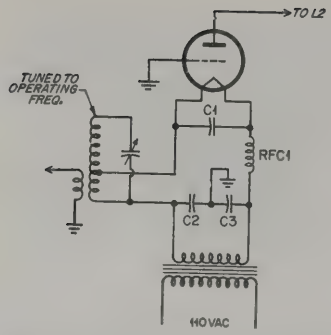
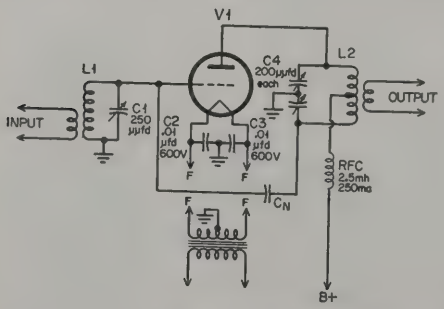


Fig. 6.4B—Variation of the input coupling method shown in Fig. 6.4-A.

It should be pointed out that the arrangement shown in fig. 6.4-A will operate satisfactorily but the stage efficiency will be improved by using a tuned circuit in the cathode. This is necessary to provide a low-impedance path from ground to cathode for the harmonic components of the tube plate current. The cathode should be tapped down on this tuned circuit for the proper impedance match. See fig. 6.4-B for this modification.

Coming now to the more familiar grounded-cathode amplifiers we find circuits that we have been using for years in class C stages. See fig. 6.4-C. There is really nothing new about the actual circuits. It is only the operating voltage on the control grid and the amount (and kind) of driving signal that are different in linear amplifiers. The conditions that go to make a good class C amplifier are the same that help to make a good linear amplifier. This would imply good tank circuit "Q", freedom from parasitics and oscillation, good mechanical lay-out and construction, etc. As outlined earlier, the stage may be operated in various modes of linear operation.

An extremely popular grounded grid circuit configuration is shown in fig. 6.4-D. This arrangement is used in the Globe Model LA-1 linear amplifier, with four EL-38's connected in parallel and operating grounded grid. Note that the screen and suppressor grids are both connected directly to ground and the high voltage (1000 volts, 300 ma) is applied only to the plates. Very stable operation is assured by this method as the amplifier is degenerative and self-oscillation cannot occur. The r-f driving voltage is capacity coupled into the low impedance cathode circuit of the amplifier and is amplified by the tubes. A regulated bias voltage is applied to the final tube grids for maximum linearity in class B. Class C operation is obtained by increasing the bias on the final grids, for cw operation.



C1—250 μμfd., variable.
C2, C3—0.01 μfd., ceramic, 600v.
C4—200 μμfd. per section split-stator, transmitting type.
Cn—Neutralizing condenser, value depends on circuit and inter-electrode tube capacity.
RFC—2.5 mh., 250 ma.
V1—Zero-bias tube, such as, 811, 805, etc.

Fig. 6.4C—Conventional grounded cathode linear amplifier stage using a zero bias tube.

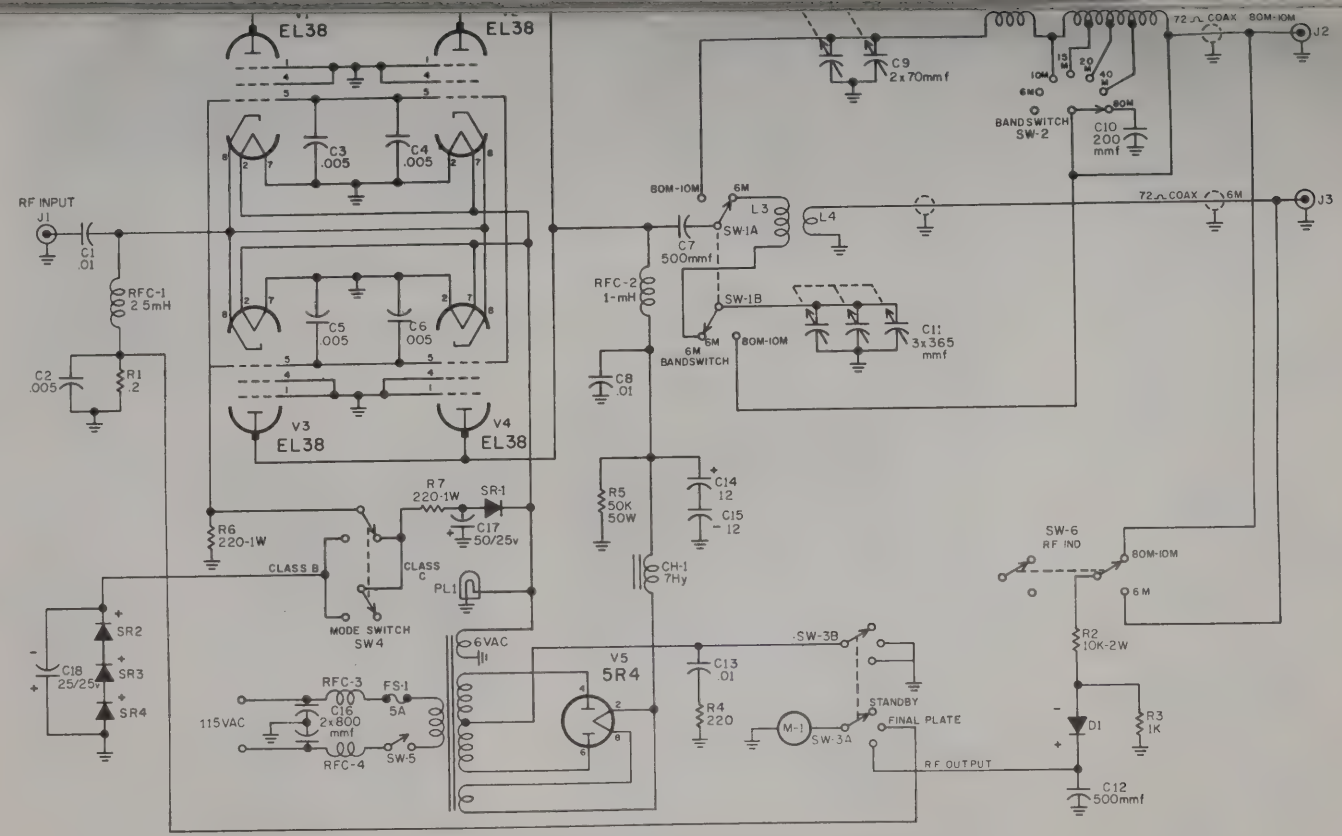


Fig. 6.4D—Globe LA-1 linear amplifier employing four tubes in grounded grid. Note the manner in which the grids are grounded.

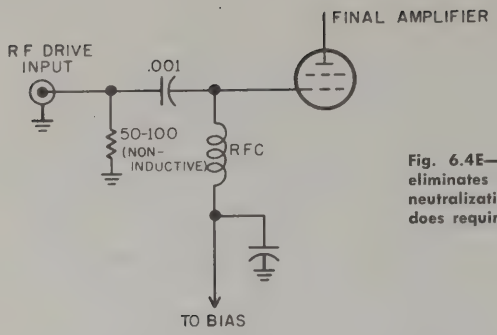
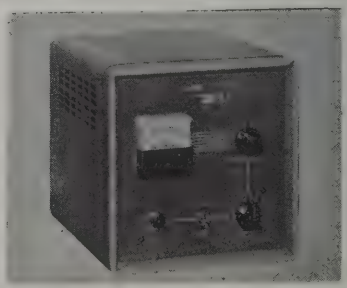


Fig. 6.4E—A passive grid circuit eliminates tuning adjustments or neutralization. This configuration does require more drive, however.

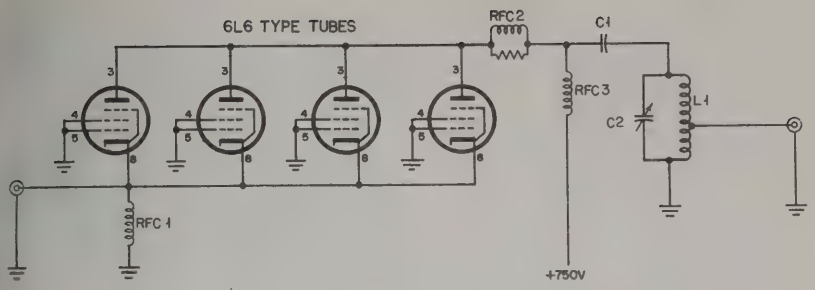
Another circuit that is gaining wide popularity in linear amplifiers is shown in fig. 6.4-E. In this circuit, a non-inductive resistor which is equal to the line impedance replaces the customary grid tank. Since the resistor is non-reactive, the stage cannot oscillate if the rest of the elements are properly by-passed. This circuit has several other advantages, in addition to stability. Because of the low grid impedance, the stage need not be neutralized. In addition, the pure resistive load presented to the exciter tends to improve its performance. This input circuit does require more drive than the grounded grid configuration, and far more than a tuned circuit would. The amount of drive can be calculated by obtaining the peak-to-peak voltage requirement for the final tube. Knowing this, the amount of power across 50 ohms, needed to develop the voltage, can be calculated by Ohms Law. This configuration is used in the Hallicrafters HT-33A in conjunction with a Penta Labs PL-172. The HT-33A will develop a full kilowatt when used in conjunction with exciters in the 50-100 watt class (output, that is).

W6GEG's popular grounded grid "pigmy rig" is shown in fig. 6.4-F. This linear, with four receiving type tubes, is capable of 150 watts peak envelope power input.

Four EL-38's connected grounded grid and operated in class B are used in the Globe linear, Model LA-1. The unit requires 7 to 15 watts of drive and will deliver 420 watts PEP power on the 80 through 6 meter bands.



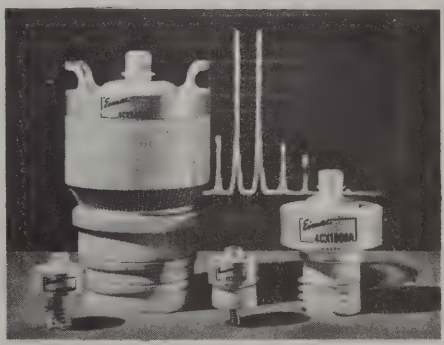
The Vantron 300-A linear amplifier uses two EL-34/6CA7 tubes and runs 150 watts PEP power input. A GZ-34 rectifier tube supplies the necessary B supply voltage.



Parts List

RFC 1, 3—2.5 mh 500 ma. RF Chokes
RFC 2—1 watt 50 ohm resistor wound end to end with No. 16 enamel covered wire
C1—500 μμfd. 10KV ceramic (TV type)
C2—Johnson 200L15 202 μμfd. .030" spacing
L1—B/W 40 JEL, for 75 meters. Tap about 4 turns from bottom

Fig. 6.4F—How much simpler can they get? This grounded grid linear was designed by Norm McLaughlin, W6GEG, and is capable of 150 watts PEP. Type 6AG7 tubes will work as well in this circuit.



Tetrode tubes, used in linear amplifier applications, produce high power gain and low distortion. The Eimac tubes shown here use a ceramic material, rather than glass, and are tailor made for class AB_1 applications.

6.5 DESIGN CONSIDERATIONS

Much of the work in developing a design procedure has been done by Reque, W2FZW, in the May '49 issue of QST magazine. In this article, he describes how to design circuits and select operating voltages when no class B tube ratings are given.

When the tube manuals are consulted, one generally finds that the audio ratings are given for class A, AB_1, and class B. It is almost possible to transfer the operation into the r-f realm—almost! In audio work, transformers are selected that properly match the tube grid impedance, or match the plate impedance to whatever load is used. One can't be quite as glib as this in r-f service. The tuned circuits must be chosen with discretion in order to accomplish the same end.

Let us take a specific example and work through it to end up with some answers that can be put to work. For the purposes of explanation let us examine the popular RCA-6146. Table 6.5-A is extracted from the RCA tube manual and illustrates how two 6146's may be used in audio service. The triode connected ratings are listed for your information. You can appreciate the advantages of pentode connected operation when you observe the maximum output ratings.

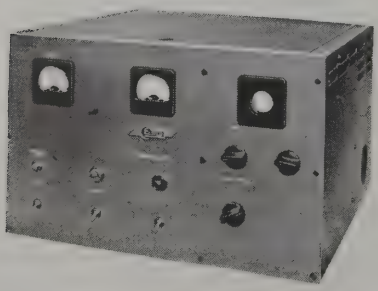
The first thing to decide is the makeup of the final product. This would include the number of tubes, push-pull or parallel operation, and class of operation. So, for the sake of the explanation, a pair of 6146's are selected to be used in some sort of linear amplifier.

ICAS Ratings	Class AB_1 Triode Con.	Class AB_1 Pentode Con.	Class AB_2 Pentode Con.
D.C. plate voltage	400 volts	750 volts	750 volts
D.C. grid #1 voltage	−100 volts	−50 volts	−45 volts
Peak AF grid-to-grid voltage	200 volts	100 volts	101 volts
D.C. grid #2 voltage	.....	200 volts	165 volts
Zero signal screen current	.....	1 ma.	0.6 ma.
Max. signal screen current	.....	27.5 ma.	21 ma.
Zero signal d.c. plate current	80 ma.	57 ma.	35 ma.
Max. signal d.c. plate current	136 ma.	227 ma.	240 ma.
Effective load (plate-to-plate)	8000 ohms	8000 ohms	8000 ohms
Max. signal driving power	0 watts	0 watts	0.03 watts
Max. signal output (approx.)	19 watts	120 watts	130 watts

Fig. 6.5A—Typical operating conditions for the popular 6146 tube in audio service. See text for conversion of figures to r-f service.

Consulting Fig. 6.5-A it can be seen that pentode connection in AB_1 will produce 120 watts output, while AB_2 operation will give only 10 watts more, or 130 watts. Since AB_2 operation involves running the tubes into the grid current region (with all the attendant troubles) it would be wise to choose class AB_1 and sacrifice the 10 watts additional output. With AB_1 operation, the design of the grid tank circuit becomes no problem at all. The only "sticker" is to furnish the grids with sufficient voltage to swing them throughout their full range, which according to the chart will require 100 volts peak-to-peak for tubes in push-pull. If the exciter is a little puny on output, you might try the following design approach: Employ link coupling between the exciter tank and the final grid circuit but use a higher L to C ratio in the exciter tank circuit. This will produce a voltage step-up as in an ordinary transformer. The resonant impedance of the grid tank circuit will be higher but this is satisfactory since the grid draws no current and appears as a very high impedance itself. The 100 volts of grid swing needed was for push-pull operation. If only one tube is used, or if two tubes are used in parallel, the grid drive needed is only 50 volts peak.

Next is the design of the plate circuit. As was stated earlier, a loaded "Q" of between 12 and 15 is very desirable. Keep this in mind, as it affects the calculations. Every tube, for a given set of operating conditions, has a plate resistance that depends on the swing of the ac plate signal voltage and current. This value need not be calculated for the manufacturers have kindly performed this task for us. In the case of the 6146 tubes in push-pull, the plate load resistance is 8000 ohms. For one tube in r-f service the proper plate load is 4000 ohms.



A pair of Eimac 4CX300's are used as amplifiers in the Eldico SSB-1000F linear. The units use a resistive input similar to that shown in Fig. 6.4-E, and a one inch oscilloscope for constant monitoring of amplifier linearity. The Eldico SSB-1000F runs 1250 watts PEP power input.

The Hallicrafters HT-33A uses a passive network with a 52 ohm load resistor similar to the one shown in Fig. 6.4E. A Penta Labs PL-172 allows the legal power input to be reached with ease.

From these values of load resistance, the desired circuit "Q", and the operating frequency, the values of tank capacity and inductance can be found. The formula is very simple and is as follows.

$$\text{Reactance (ohms)} = \frac{\text{plate load resistance}}{\text{loaded circuit "Q"}}$$

The reactance is the inductive reactance or capacitive reactance that the coil and capacitor will have at resonance. At resonance, both reactances will be equal. By substituting the known values in the above formula it will be found that a reactance of 533 ohms will be required. The exact values for the coil and capacitor may now be found by solving the following formula:

$$\text{C in (mfd.)} = \frac{1}{6.28 \times F \times X_c}$$

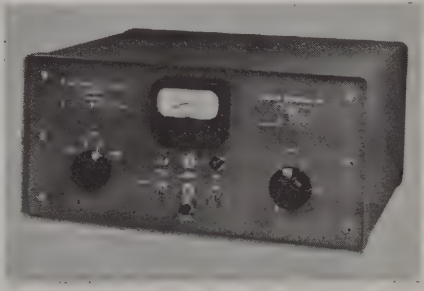
$$\text{L (in microhenries)} = \frac{X_L}{6.28 \times F}$$

F = frequency in mc. X_c and X_L = 533 ohms (previously determined)

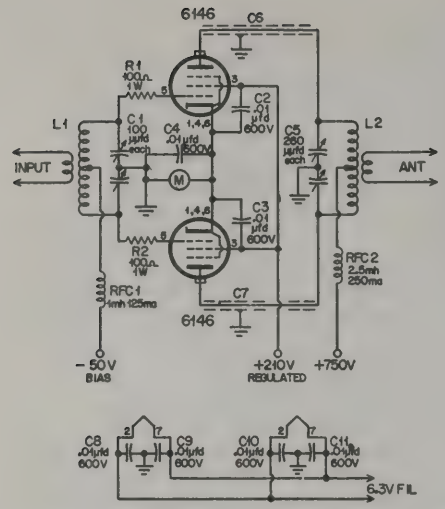
Solving these two formula, for the 75 meter band, the capacity will equal .000075 mfd (or 75 mmfd) and the inductance required will be 21.2 microhenries. The effective capacity required is 75 mmfd., so that in a push-pull circuit a split stator 150 mmfd. per section variable capacitor should be used. To cover the entire band, a practical value would be somewhat more than 200 mmfd. per section.

Someone is bound to be thinking "Why not use the tubes in parallel and avoid the use of a split stator capacitor and center-tapped coil?". The idea is sound, but the appropriate changes must be made in the tank L and C values. Since the plate load resistance of a single 6146 tube is 4000 ohms, the two tubes connected in parallel will yield 2000 ohms. When this value is inserted in the above formula, you will note that the tank circuit capacity has increased by a factor of four (300 mmfd., rather than 75 mmfd.) for parallel operation.

The size of the tank coil has also been changed because of parallel operation. It is now ¼ of its former self, and has an inductance of only 5.2 microhenries.



The popular 813 tube is used in the Central Electronics Model 600L linear amplifier, and operates in class AB_2. With 8 watts of drive, this unit will deliver 400 watts PEP power on all bands between 160 and 10 meters. The 600L uses band-pass tuned circuits to avoid tuning adjustments. An 812A triode is used to regulate the screen grid voltage.



C1—100 μμfd., per section split-stator.
C2, C3, C4, C8, C9, C10 C11—0.01 μfd., 600v., ceramic.
C5—260 μμfd., per section split-stator.
C6, C7—Coaxial condensers made of RG-58/U or RG-59/U with the shield grounds.
M—0-300 ma., d-c milliammeter.
R1, R2—100 ohms, 1w.
RFC1—1.0 mh., 125 ma., r-f choke.
RFC2—2.5 mh., 250 ma., r-f choke.

Fig. 6.5B—Push-pull arrangement for class AB$_1$ using 6146 tubes. See Fig. 6.5D for the coil details.

This means that if you are in the habit of buying commercial wound coils, for parallel operation you should purchase a 40 meter coil for use on 80 meters, and a 20 meter coil for use on 40 meters. Even in the push-pull amplifier, turns must be removed from a standard 80 meter coil in order to attain the proper L to C ratio.

And so you must make your decision as to which circuit you will use—push-pull or parallel. It is a matter of personal preference, and as the man says, "You pays your money and you takes your chances", or something like that.

This procedure of using the proper L to C ratio is very important for proper operation of linear amplifiers. If the basic rules are followed the tubes will run cooler, the maximum power will be transferred to the antenna, the harmonic out-

C1—100 μμfd. single section air variable (Cardwell ZR-100-AS)
C2—0.001 μμfd., 500v. mica
C3, C4, C5, C6—0.01 μfd., 600v. disc ceramic
C7—0.001 μfd., 1000v. mica
C8—Cardwell MR-260-BD dual section transmitting condenser with both sections tied in parallel
R1, R2—100, 1w. carbon
R3, R4—47 ohms, 1w. carbon
M—0-300ma. DC milliampmeter

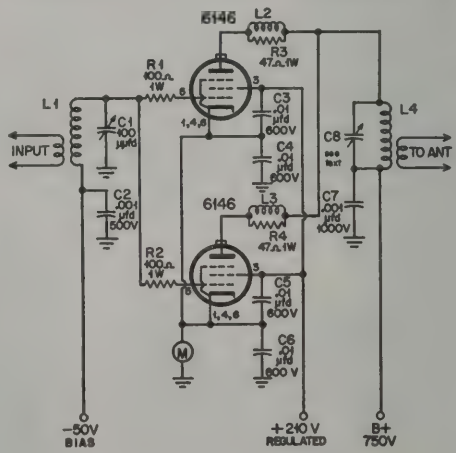


Fig. 6.5C—A parallel connected 6146 class AB$_1$ amplifier capable of 120 watts output. Note the difference in the plate tank L/C ratio when compared to Fig. 6.5B. Coil information is supplied in Fig. 6.5E.

	80 Meters	40 Meters	20 Meters	15 Meters	10 Meters
L1 Grid Tank	B&W 80 MCL	B&W 40 MCL	B&W 20 MCL	B&W 15 MCL	B&W 10 MCL
L2 Plate Tank	B&W 80 BVL	B&W 40 BVL	B&W 20 BVL	B&W 15 BVL	B&W 10 BVL

Note: Turns are removed from the BVL coils until capacitor C5 resonates at the high end of the band, with the plates ⅓ meshed, except on 80 meters where it should be ⅔ meshed.

Fig. 6.5D—Coil details for fig. 6.5B.

put will be less, in other words things will be running more efficiently.

Fig. 6.5-B and fig. 6.5-C are operational circuits for the two amplifiers designed in the text. The values were determined from the formulas and on the basis of good engineering practice.

	80 Meters	40 Meters	20 Meters	15 Meters	10 Meters
L1 Grid Tank, End Link	B&W 80 MEL	B&W 40 MEL	B&W 20 MEL	B&W 15 MEL	B&W 10 MEL
L2, L3 Parasitic Chokes	10 turns, #22 enam. wire wound on R3 and R4				
L4 Plate Tank, End Link	B&W 40 BEL	B&W 20 BEL	B&W 15 BEL	B&W 10 BEL	B&W 10 BEL

Fig. 6.5E—Coil details for fig. 6.5C.

6.6 TWO TONE TESTS AND THEIR MEANING

The "build it yourself" hams often speak of "two tone tests." They are referring to the simple test that can be performed on a SSB transmitting system to check the linearity of the amplifiers. It is necessary to have a "yardstick" with which to check the amplifiers that have been so carefully designed and constructed.

Briefly this is the idea of the two tone test: Feed two steady sine wave audio signals into the SSB exciter input. You can use two audio oscillators, or inject carrier and feed just one tone into the microphone input. The amplitude of these two signals should be kept equal for the tests. Connect an oscilloscope to the output of the last amplifier to be checked. The method is described in Chapter IV. The dummy load should approximate what the amplifier "sees" when it looks into the transmission line (purely resistive, it is hoped). Set the oscilloscope sweep, the carrier level and modulation level until a trace similar to fig. 6.6-A is obtained.

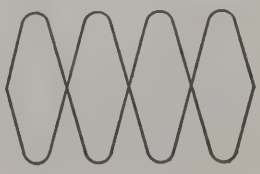


Fig. 6.6A—Oscilloscope pattern showing the ideal envelope with a two-tone test in progress. No serious non-linearity present here.

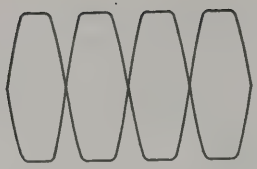


Fig. 6.6B—In this two-tone test pattern there is peak flattening due to either excessive drive, poor regulation of the driver stage, or insufficient antenna coupling.

This is the way the pattern will look if everything is operating properly. Increase the level of the two tones until the system starts to overload—this will be evident by flattening of the peaks as shown in fig. 6.6-B. If this point of maximum linear input is considered less than the final amplifier is capable of, some checking is in order. Make sure that the loading on the final is heavy enough and that the driver stage is not overloading before the final reaches its maximum grid driving requirements. A quick test for insufficient final plate tank loading is to detune the final slightly off resonance, and watch the pattern. If the flattening disappears as the oscilloscope pattern decreases slightly, the final is not loaded heavily enough. Steps must be taken to increase the antenna coupling. However, if the flattening remains, the trouble is in the driver or the coupling arrangement between the driver and final. Slightly tighter coupling between the stages, or at least more efficient coupling, must be accomplished.

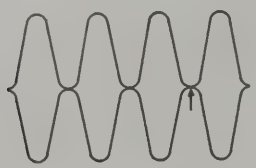


Fig. 6.6C—Sloppy cross over characteristics are seen in this two-tone test pattern. The grid bias should be reduced until the cross-over (shown at the arrow) is a sharp "X" as seen in fig. 6.6-A.

If the oscilloscope pattern looks like fig. 6.6-C, the trouble is something else. The grid bias is too high and must be reduced until the two sine wave patterns cross the center line with a perfectly sharp "X". An "X" marks the spot, to make a very poor joke! This type of distortion is present at all levels of signal input and has been nicknamed "cross-over distortion". Reducing the gain when operating the transmitter on the air will do little to clean up this latter type of distortion. Gain reduction would however, help the peak flattening distortion mentioned first. Don't reduce your operating bias to the point where the tubes are dissipating more than their ratings call for.

Power Ratings

A so-called "hi-fi" audio amplifier is another form of linear amplifying device. These units are power rated at so many watts output, and at a specified distortion level. A typical hi-fi amplifier data sheet might read: Harmonic distortion, 5% at 32 watts- Intermodulation distortion, 1.5% at maximum rated power. A single sideband exciter or linear amplifier is rated in a similar manner but the manufacturers often "forget" to mention how much distortion is present at the maximum power output level. If no figure for distortion is given, then one must assume that it is within the acceptable limits for such a system. The distortion products will increase very rapidly as the overload point is approached so the published ratings will be reasonably accurate. The standard method of rating sideband exciters and

linear amplifiers is in terms of "peak envelope power", or PEP for short. This can be defined as the rms power developed during the peak r-f cycle. Peak envelope power can be found by measuring the amount of r-f voltage developed across the dummy load resistor, with a high frequency ac voltmeter, during a two tone test. The level of the two tones should be just below the point that produces overload and noticeable distortion products. The voltage squared, divided by the resistance of the dummy load will reveal the peak envelope power. However, since few hams have r-f voltmeters (Hewlett Packard, GR, etc.) it is necessary to make the measurements with an oscilloscope. The deflection plate is connected to the dummy load resistor through a high voltage capacitor to avoid short circuiting the cathode ray tube centering voltage. A voltage step-up device (such as a tuned circuit) should not be used unless the exact step-up ratio is known. Since the oscilloscope displays peak-to-peak voltage, the voltage represented by the total amplitude must be divided by two. This, then, is the peak envelope voltage. But the formula is based on rms values so the peak voltage must be multiplied by .707 to obtain the rms voltage suitable for the formula. This final figure is "E" and is substituted in the power formula (PEP- E^2/R), to establish the peak envelope power.

How does the kilowatt crowd insure that they are not exceeding the maximum power input? Is it necessary to keep an oscilloscope and dummy load on the transmitter at all times? Obviously the answer is no. The FCC has provided the Amateurs with a very liberal ruling to cover the situation. The power input is determined in the usual fashion, i.e. plate voltage times plate current to the final amplifier. When modulating the operator is not allowed to "kick" the plate current meter beyond the point where the legal power limit is exceeded. In other words, if the final has 2000 volts on the plate, the loudest voice peak should not push the meter beyond 500 ma. I can see visions of spring loaded and damped plate current meters dancing through the minds of the reader. As usual, the FCC is always one jump ahead and have specified that the meters shall have a time constant that does not exceed 0.25 seconds. This would lead one to believe that the indicator must reach 63% of its reading in less than 0.25 seconds. This figure is easily obtained in commercial meters with the customary amount of damping.

Since the meter cannot be expected to follow the rapid excursions caused by voice modulation, how much power input are we actually running when the meter indicates a kilowatt of input? The true power input will vary depending on the type of modulation but it is safe to say that the power input will be more like two kilowatts. Several manufacturers have taken advantage of this fact and rate their amplifiers at two kw peak envelope power input, and it's all perfectly legal.

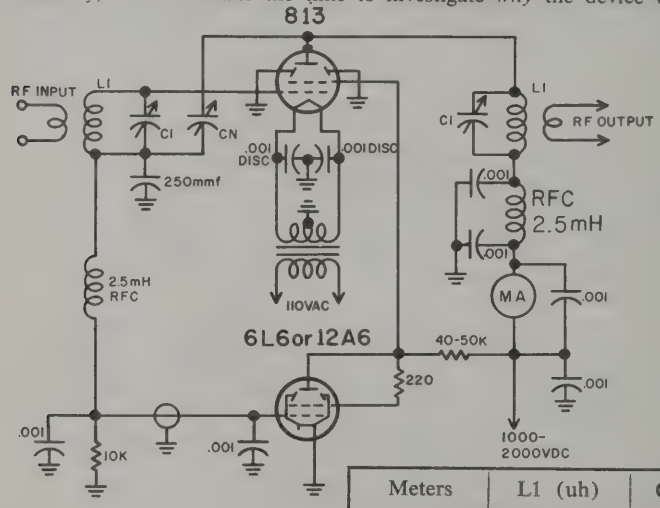
Power In Versus Power Out

As you may have gathered from the earlier explanation of peak envelope power, it is somewhat difficult to measure the power output. It does require a test equipment set-up and there are always possible sources of error. Fortunately there is a correlation between the plate current reading and the PEP output. The stage efficiency plays an important part in calculating the PEP output, as you can well imagine. Depending on this factor, the PEP input may be anywhere between 1.4 and 2 times the PEP output. There are other factors too, but it is safe to assume that the PEP input will be approximately 1.57 times the dc input, when measured with a two tone test signal. Unless these linear amplifier tests are made below the point of distortion you are simply kidding yourself. When peak compression or flattening occurs the power output will appear to be higher. But the additional power will be nothing but distortion and will add nothing in the way of useful information to your signal.

6.7 A CLASS C LINEAR R-F AMPLIFIER

As soon as the eyebrows stop clicking and the unbridled laughter ceases, clear your mind of all the facts that you have just been reading, about classes of operation. Hold off regarding opinions on my sanity for a few pages; I must warn you, I have a lawyer. Strange as it may seem, I am about to describe an amplifier that uses no grid bias supplies (either regulated or otherwise), obtains its bias by rectifing the driving SSB signal, and does not employ screen regulation! As a matter of fact, the screen voltage is obtained through a huge dropping resistor, and the voltage kicks around wildly!

Possibly some background information will help to soften the blow. Some time ago, the author did a conversion on the Heath DX-100 to double sideband (April 57 CQ Mag.). In this conversion, the grid voltage was not stabilized, the screen voltage was not regulated, and the original clamp tube was left in the screen circuit. Many of the more learned hams scoffed at such an idiotic conversion, but the fact remained that the 6146 final in the DX-100 *did* operate as a linear! The r.f. peaks were extremely sharp and the cross-over "X" was just like the book said it should be. Unfortunately, I did not take the time to investigate *why* the device worked.



Values for L1 and C1 are shown in the accompanying chart.

Meters	L1 (uh)	C1 (uuf)
75	7	270
40	3.72	145
20	2.34	75
15	1.34	50
10	.95	36

Fig. 6.7A—Schematic for the ZL linear (ZL1AAX). Note that no external bias supply is used and the screen voltage is not regulated. L1 should resonate on the desired band with C1.

Some time before this, Lester A. Earnshaw, ZL1AAX, was conducting experiments along the same line, that is, using a clamp tube in conjunction a class C amplifying stage. Lester perfected the circuit, and it is this design that is presented here (fig. 6.7-A). His design has been duplicated with the same success by hams

all over the world, but the comment "I don't believe it" still persists whenever it is described over the air. The whole idea is akin to the bumble bee which, according to the laws of aerodynamics is too heavy in the body and too small in the wing to be able to fly. Fortunately for the bee kingdom, they do not know anything about aerodynamics!

The circuit for the ZL Linear (ZL1AAX) uses the popular 813, in a class C circuit, but with a clamp tube connected from screen to ground. According to the tubes manual, the 813 requires a 10K grid leak and 40K screen resistor, at 1500 vdc. The action of the clamp tube is very interesting. In the absence of drive there will be no rectification in the grid circuit of the 813. Since there is no voltage developed across the grid leak, there will be no bias applied to the clamp tube, and it will draw a large amount of current through the 40K dropping resistor. That will produce exactly the same effect as if a low value resistor were connected between screen and ground. There will be a large voltage drop across the screen resistor and *low voltage on the screen*. All of which adds up to low plate current in the 813, in the absence of drive. Now, observe what happens when a small amount of r-f is fed to the 813. There will be some rectification at the grid, voltage developed across the grid leak (and supplied to the clamp, for bias), and less current flow through the clamp tube. This means that the latter will partially "unclamp", thus allowing the 813 plate current to increase. *The increase will be in proportion to the grid drive applied.* This will be true so long as the clamp tube begins unclamping the moment negative bias is applied to its grid. If it takes say, 2 volts of negative bias before the tube begins to raise it's plate resistance, then the clamping action will not be linear.

A somewhat similar circuit is used in the popular G2MA linear, but in his circuit, a separate diode does what the grid of the 813 does in the ZL linear. In tests made by Lester, ZL1AAX, the performance of the ZL linear and the G2MA linear are nearly identical. Peak flattening takes at about the same level and the output is similar with either circuit configuration. Basic differences between the ZL and G2MA linears is shown in figs. 6.7-A and 6.7-B.

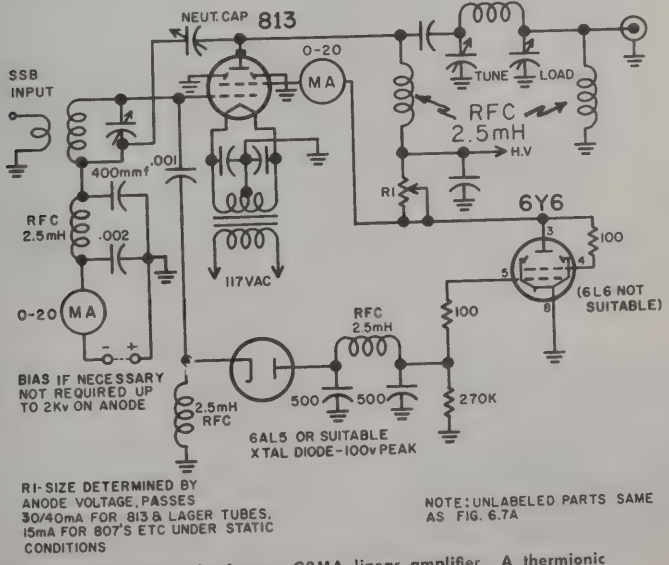


Fig. 6.7B—The famous G2MA linear amplifier. A thermionic diode supplies the clamp tube voltage.

The loading effect of the amplifier upon the driver stage is exactly the same as that of a zero bias class B linear. In this class (unlike class AB_2) grid current is drawn for all amounts of input. Thus, the impedance is normally low, and varies lower. In AB_2, when grid current is not drawn, the impedance is high until the moment of grid current flow, then the impedance drops considerably. Of the two, zero bias class B offers the least variation in grid impedance and therefore the demand upon the exciter regulation is less stringent. The ZL linear behaves exactly like a zero bias class B linear and the drive requirements are the same.

At this point, a word about tubes is in order. ZL1AAX has experimented with a pair of 6146's and found the combination unsatisfactory when clamped with a 6AQ5. The 6AQ5 appears to have insufficient clamping action, in this application, when connected as a triode. Many have had great success with other tubes; John, ZL2AG has the system working nicely with a 4-125A, several ZL3 chaps have had success with the 807, and an unknown VK is reported to be using a Philips equivalent of the 4-65A. There is no apparent reason why the system cannot be applied to almost any tetrode or pentode.

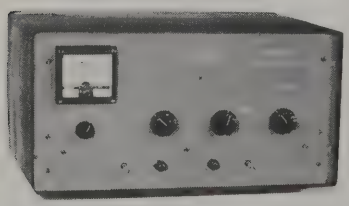
The choice of clamp tube is somewhat more critical. Earlier it was stated that the increase in the 813 plate current will be in proportion to the input *if* the clamp begins unclamping the moment bias is applied. Many tubes have this property. The 12A6 and 6F6 are excellent tubes, as is the 6L6 and 5881. The usual 6Y6 is unsatisfactory for the clamping action is far too severe. No doubt there are other tubes that will have the linear action required.

Technical Considerations

Excessive grid drive will cause too much grid current and will drive the 813 into the class C region. This should be avoided like the plague for someone (probably another ham) is sure to saw your tower in half. With 1100 volts on the 813, 2.5 mils of grid current, (on a two tone test) seems to be about maximum. As the plate voltage is increased, the grid drive requirement will decrease.

It is quite important that the correct LC ratio be used in the plate tank circuit. There should be more capacity than would be used in a class C stage. If the reader has any doubts about this he should refer to the formulas for L and C given earlier in this chapter. In addition, the amplifier must be completely stable. It must be checked without drive by varying the input and output capacitors, across resonance. The smallest trace of grid current or plate current fluctuation must be eliminated. Neutralization is a necessity.

Rough tests of stage efficiency indicates that the ZL linear has an edge over the zero bias class B linear stage, which is usually thought of as the most efficient type of linear amplifier. The ZL linear should gain wide popularity because of the lack of regulated power sources, ease of adjustment, and high efficiency. The fact that it is so similar to existing transmitter circuits, should make it a natural for converting these transmitters to single sideband.



The Lakeshore P-400-GG uses four 6CN6 tubes in a grounded grid configuration and operates on 80 through 10 meters. With a driving signal of approximately 20 watts, this unit will deliver 400 watts PEP to a 52-300 ohm load.

The G2MA Linear

Another form of clamp tube linear was conceived and designed by David D. Marshall, G2MA. The circuitry varies slightly from the ZL1AAX linear that has just been described. The main difference is the method of obtaining bias for the clamp tube. You will note from fig. 6-7-B that the G2MA linear uses no grid bias and no grid bias resistor. There is no load resistor for voltage to develop across and clamp tube bias is obtained by rectifying the SSB energy in a thermionic diode.

Grid bias may be necessary, however, when more than 2000 volts is used on the final plate. As an example, the 813 will operate at 2000 volt with no bias and exhibit a static plate current of 25- 30 ma. With 3000 volts on the plate (sure, the 813 will take it) about 12 volts of negative grid voltage will be required to produce the same plate current. A 4-400 at 3000 volts will require about 25 volts of bias.

It is very desirable to meter the amplifier screen current to make sure that under maximum drive conditions the current never rises above about ¼ of the bleeder current through the dropping resistor and clamp tube. The clamp tube should always be reducing current faster than the screen of the final amplifier can "take it". Thus the total current through the screen dropping resistor will always be reduced when bias is applied to the clamp tube. This in turn, raises voltage and causes the final to draw more current.

As David says "It all sounds very complicated but it really is not. If the constructor knows 'how many beans make five', he should experience no difficulty in obtaining proper operation in a few minutes. Oscilloscope patterns show the linearity to be as good as when the same tubes are operated in class AB_1 or AB_2. Since current is drawn over the entire r.f. cycle, grid swamping seems to be unnecessary".

A Gated Linear Amplifier

There is still another form of screen controlled linear amplifier that will be of particular interest to hams in the "grounded grid school". The circuit shown in fig. 6.7-C was designed by "Norm", W6EDD. Two of the popular war surplus RK-65's

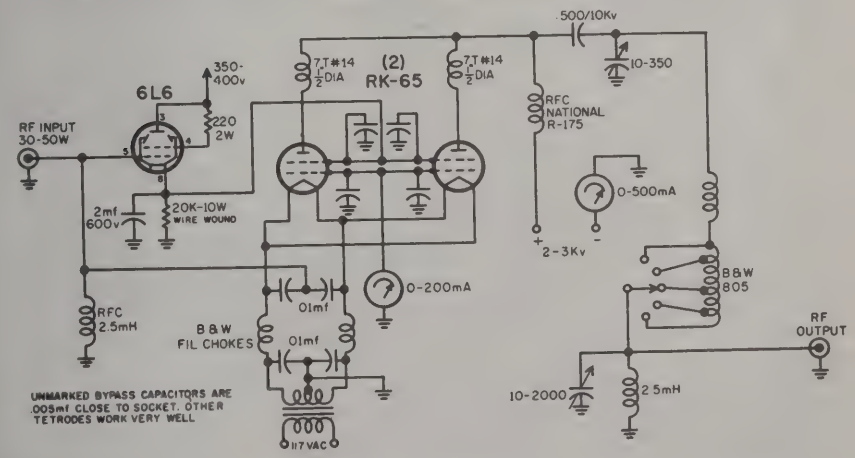
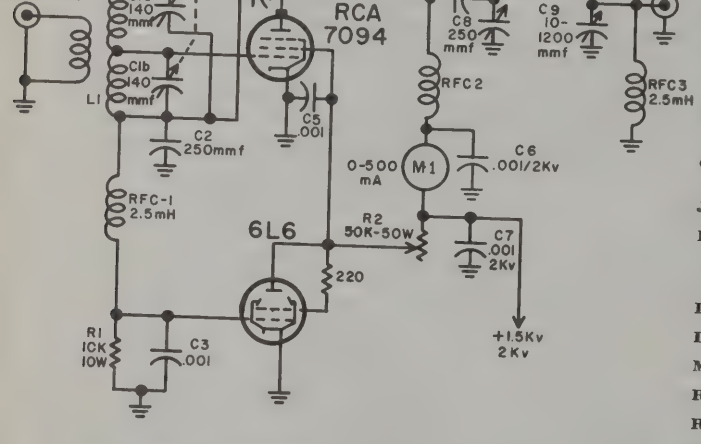


Fig. 6-7C—A grounded grid gated linear amplifier, designed by Norm Roller, W6EDD. The 2 mfd. capacitor on the 6L6 cathode must be an oil-filled type and is used to remove high peaks and improve 3rd order harmonic distortion. The .005 mfd. screen and control grid bypass capacitors should have very short leads and be mounted near the tube socket. The 350- 400 volt supply should have 10- 15 percent regulation. If taken from high voltage bleeder, connection to plate of the gating tube should be regulated.

are used in a grounded grid configuration. The control tube (a 6L6) is connected in series with the screen and the 380 volt power supply. In effect, the 6L6 acts a series gate tube rather than clamping the screen voltage to ground as is the case with the ZL and G2MA linears.

All the advantages of grounded grid operation are retained with this circuit, and far less grid drive is required to "get the show on the road". R-F drive from the exciter is applied to the 6L6 and the heater circuit of the RK-65's. The positive r-f cycles cause the 6L6 to draw more current, thereby increasing the screen voltage. A B&W filament choke eliminates the possibility of r-f getting back into the power line. A standard pi-network output circuit effectively matches the final stage to the antenna. The RK-65's will handle 2 kilowatts PEP input with ease.



C1a, C1b—Dual 140 mmfd variable capacitor (Hammarlund HFD-140)
C2 — 250 mmfd., 600 volt mica
C3, C5—.001 mfd. disc ceramic
C4—1-8 mmfd. neutralizing capacitor
C6, C7, C10—.001 mfd., 2 kv
C8—250 mmfd. variable capacitor (E. F. Johnson type 250E20)
C9—10-1200 mmfd., 4 section broadcast capacitor (J. W. Miller #2104)
J1, J2—UHF style connectors (Amphenol)
L1—29 turns, #24, 1" dia., 1⅞" long, centertapped (B&W 3015 or Air Dux 816T) Link: 4 turns, #22 plastic covered wire.
L2—6 turns, #14, 1" diameter, 1¼" long.
L3 — Rotary inductance (E. F. Johnson 229-201)
M1—0-500 ma. plate current meter
R1—10K, 4 watts (two 2 watt carbon 20K)
R2 — 50K, 50 watt adjustable wire wound resistor
RFC1, RFC3—2.5 mh. radio frequency choke (J. W. Miller #6302)
RFC2—Plate choke (Rapar RL-102)

Fig. 6.7D—An "Americanized" version of the ZL linear.

6.8 LINEAR AMPLIFIERS USING SURPLUS TUBES

The 304TL and Western Electric 701A are plentiful and inexpensive, on the war surplus market. Unfortunately, both these tubes have extremely high interelectrode capacities which reduce their usefulness on the 10 and 15 meter band. However, they do represent an inexpensive way to get a kilowatt going on the lower frequency bands.

Two units are described here, one using a single 304TL and the other incorporating a pair of WE-701A's. The 304TL amplifier is shown in fig. 6.8-A and is conventional in most respects. The filament was wired for 10 volt operation because of the availability of 10 volt transformers, and the desire to avoid copper buss for filament wires. The 304TL filament requirements are 10 volts at 12.5 amperes or 5 volts at 25 amperes. The center-tap of the filament is used as the cathode return point. Both sides of the filament voltage line are bypassed to ground (C5-C6) at the tube socket with good quality mica capacitors. The cathode current meter is connected from the filament center-tap to ground, and C7 grounds the meter for r-f. The parasitic resistors R1 and R2 are in the grid circuit and the grid neutralizing

R1—27 ohms, 2w.
R2—23 ohms, 4w (2 parallel 47-ohm 2w.
R3—5000 ohm, 10w. non-inductive.
R4—25,000 ohm, 50w, with slider.
C1—250 μμfd., single section, air.
C2, C4, C7—0.01 μfd., 600v., disc ceramic.
C3a, C3b—dual, 40 μfd., 450v., electrolytic (insulated from chassis).
C5, C6—0.001 μfd., 600v., mica.
C8, C9, C10—50 μμfd., 5000v., vacuum padder (used only on 80 meters).
C11—150 μμfd., per section, split-stator transmitting, air, variable.
C12—0.001 μfd., 5000v., mica.
C13—600v mica; for 80 meters—1200 μμfd., for 40 meters—600 μμfd., for 20 meters—300 μμfd.
C14—16 μfd., 450v., electrolytic.
CN—neutralizing condenser, 12 μμfd., maximum capacity.
RFC1—1.0 mh., 500 ma. r-f choke.
CH1—10 h., 90 ma., filter choke.
M1—0-15 milliammeter.
M2—0-500 milliammeter.
T1—350-0-350v at 90 ma., 5.0v at 3 amp.

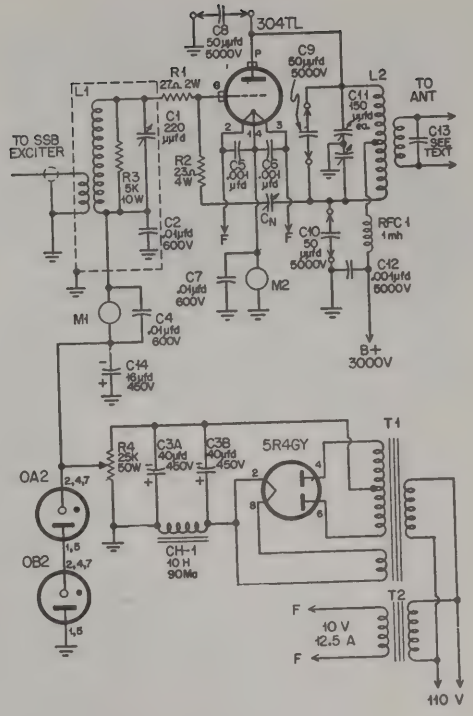


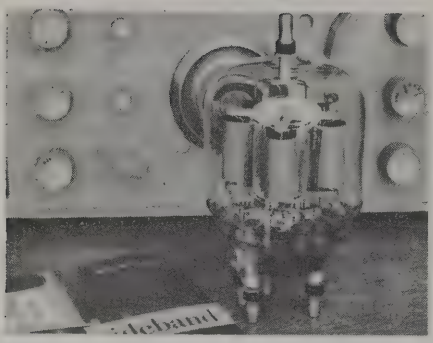
Fig. 6.8A—Kilowatt linear amplifier using the surplus 304TL.

leads respectively. These values are not critical, but the use of non-inductive resistors is mandatory.

The 23 ohm resistor (R2) should be at least a four watt unit. During tune up (with light antenna loading) the circulating r-f currents in the tank and neutralization circuit can run very high. This condition does not have to exist long before resistor R2 goes up in smoke. The resistors are essential to the operation of the amplifier. Without them, the amplifier will "take off" and go over the hill before you can even think about hitting the plate voltage switch.

The plate tank circuit is conventional with the possible exception of the vacuum capacitors (C8, C9, C10). They are necessary to enable the B&W 40TVH coil to resonate over the entire 3.5-4.0 mc band. It was necessary to use two 50 mmfd capacitors connected "split stator"— that is, one from each end of the coil to ground and a third one across the entire coil. Connected as shown, it is possible to cover the entire 3.5- 4.0 mc band with the 150 mmfd tank capacitor. The 50 mmfd vacuum capacitors are the 5 kv units found in the war surplus antenna switching units for the ARC-5 series transmitters.

The output link coil (fig. 6.8-A) has a fixed mica capacitor, C13, connected across it, and was found necessary in order to tune the link to resonance at the operating frequency. Even though the transmission line is flat, difficulty can be encountered in properly loading the final. An ordinary 600 volt postage stamp mica capacitor works fine. One precaution, however, never operate the amplifier



Thanks to the U.S. Navy, you can buy a pair of tubes for a kilowatt final for less than four dollars. The 701A is a radar pulse pentode that works very nicely in linear amplifier service at frequencies up to 15 mc.

with the antenna or dummy load disconnected. The capacitor across the link will pop before you can say Marconi! The same results can be accomplished by inserting a variable capacitor in series with the ground end of the link, as shown in fig. 6.8-B. About 50- 100 mmfd capacity would be necessary to resonate the link.

A check of the output power at 4.0 mc with a "gallon" input indicates a measured power output of 550 watts into a dummy load. This is an efficiency of 55% and is well below the predicted 65% for this type of service. However the tank "Q" is loaded rather high, which probably accounts for the lower output power. No grid current will be drawn by the 304TL until the 750 watt input point is exceeded.

The driving power required is approximately 15- 20 watts PEP. A swamping resistor of higher value could be substituted for R3 if less driving power is available. However, it is still a good idea to generate more power than you need, and swamp out the excess.

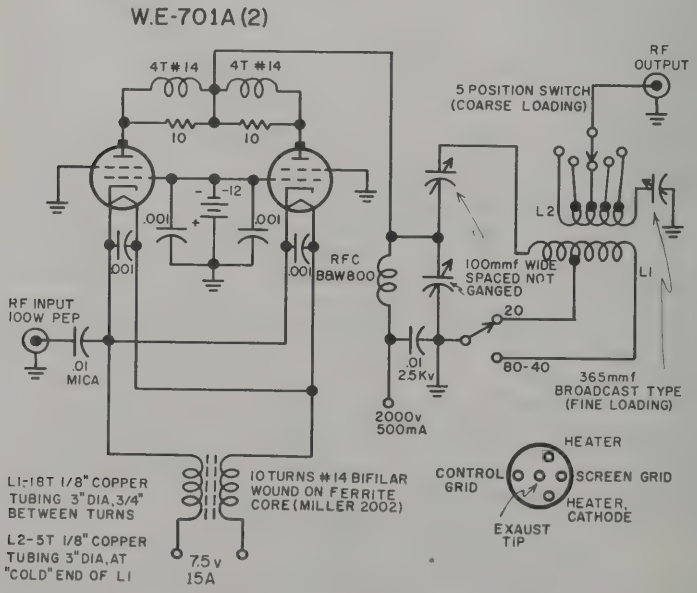
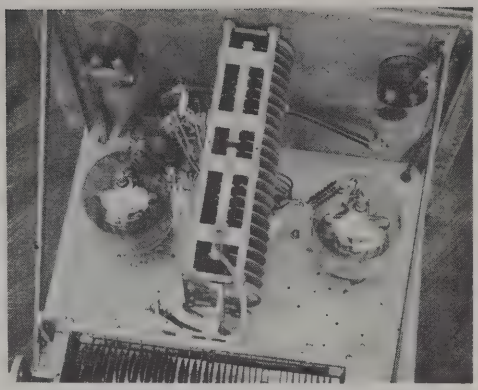


Fig. 6.8B—Linear amplifier using surplus 701A tubes.



Business end view of John Harmon's (W6RDJ) 701A linear amplifier. A trick circuit is used by John to overcome the high output capacity of these jugs.

A 701A Amplifier

The Western Electric 701A is a radar pulse amplifier tube and has been offered on the surplus market for as low as $1.95. Although the inter-element capacity is very high, it will work very well on 20 meters. By using the "trick" circuit shown in fig. 6.8-B, a *single tube* can be made to work on the 15 meter band. The capacity presented by two tubes is excessive for proper operation on 15 meters, however. The circuit shown in fig. 6.8-B was designed and constructed by John Harmon, W6RDJ. The two 701A's are operated in grounded grid fashion. Note that the screens are grounded directly, while the control grids are bypassed for r-f but supplied with a -12 volts dc. This negative bias is required to keep the idling plate current within the 200 watt plate dissipation rating. The r-f in the cathode circuit is isolated with a bifilar wound filament choke. Originally, the core for this choke was the ferrite rod from a transistor radio antenna, Miller #2002. The wire was stripped off and rewound with #14 wire. Each filament is bypassed at the socket, and the r-f is supplied to the heater pin that is internally connected to the cathode. Be sure to observe these connections or hum modulation will appear on your signal.

The output tank circuit is somewhat unusual in that it is series tuned. John, W6RDJ, claims that this increases the efficiency on the higher frequency bands. On 20 meters the upper 100 mmfd. capacitor is used as the tuning element and the lower 100 mmfd. unit is left at minimum capacity. On the 40 and 80 meter bands, additional turns are switched into the circuit and both capacitors are adjusted for maximum output. The plate impedance is approximately 2250 ohms per tube.

A socket can be made for the 701A tube by clamping it to an insulated plate that has been appropriately drilled to accept the tube pins. Cut slits down the side of a two inch length of copper tubing so that it will slide over the pins. Socket connections for the 701A are also shown in fig. 6.8-B.

With 2000 volts applied to the plates, adjust the number of battery cells in the control grid circuit for a static plate current of 75 ma. This will keep the 701A's within their dissipation rating of approximately 200 watts. When the amplifier is driven, the plate current should rise to approximately 400 ma on peaks.

6.9 A LINEAR FOR THE HAM WHO HAS EVERYTHING

Like many-another budding DX'er, the author has pined for a "linear to end all linears". As you no doubt know, the ego can be permanently damaged by holding down first place—at the bottom of the pile-ups! My deluxe dream amplifier was mentally pictured as having everything but the kitchen sink (no water cooling please)

and would handle the legal maximum without developing anodeitis. Naturally, all the features I wanted in the linear could not be obtained commercially (no, not even with time payments). A few requirements were kept in mind while the design of "my dream linear" materialized. They were:

1. Operation in Class AB_1 for minimum distortion consistent with reasonable efficiency.
2. Operate at the legal limit without requiring more than 2.5 kv (I only have one supply).
3. An input circuit that would be compatible with the popular 6146 exciters (HT-32, KWM-1, etc.) and would "soak up" the 30-50 watts of drive without requiring matching pads or attenuators.
4. Excellent stability without requiring (ugh!) neutralization.
5. Fully metered circuits (no more meter switching!).

The tube selected for the job was the Penta Labs type PL-172. This "jug" contains several novel design features that put it in a class all by itself. It is a pentode, has a 1000 watt plate dissipation rating, and is extremely compact. Compare the PL-172 photograph with those 304TL's you are using! The grid to plate capacity is extremely low which allows high stage gain and stable operation. Actually, if a grid tank and suitable neutralization were used, the PL-172 could be driven into the illegal region with a 10A! The perveance of the tube is so high, that the legal limit can be reached with less than 2000 volts on the plate! The tube uses an indirectly-heated cathode with a heater rating of 6.0 volts at 7.5 amperes. It will plug into a standard 829/832 socket or a commercial socket (PL-184) is available.

The circuit for the deluxe amplifier is shown in fig. 6.9-A. The components inside the dotted lines are contained in the shielded box that can be seen in the photographs. R-F energy from the driver is applied to a "dummy load" resistor. In effect, the source "looks into" a purely resistive load and changes in grid impedance (such as when the grid draws current) will have little effect on the driver. It is the voltage developed across R1 that swings the grid through its cycle. The

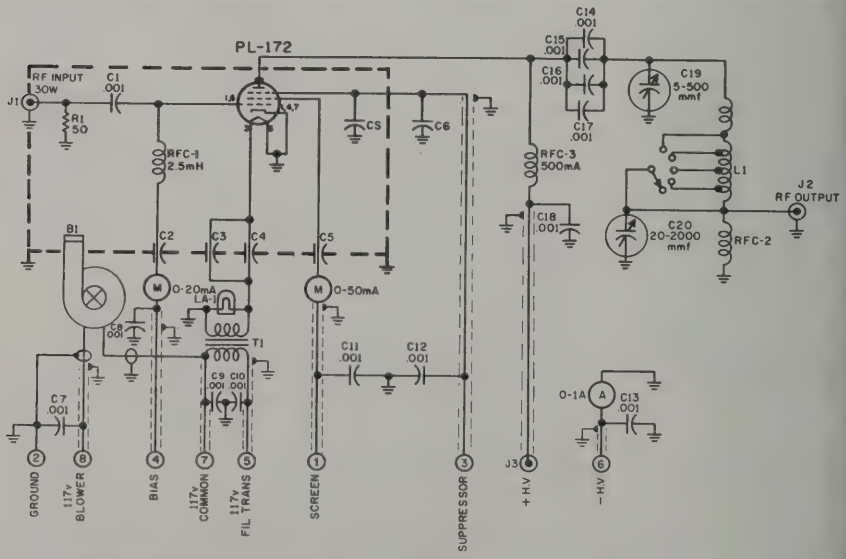


Fig. 6.9A—The deluxe PL-172 pentode linear. It will operate class "C", if desired.

TYPICAL OPERATION — Class-AB$_1$ Linear R-F Amplifier[1]
Single-Sideband, Suppressed Carrier

D-C Plate Voltage	2000	2000	2000	2500	2500	3000	volts
D-C Screen-Grid Voltage	350	350	500	500	500	500	volts
D-C Suppressor Grid Voltage	0	75	75	0	75	75	volts
D-C Control-Grid Voltage[2]	−70	−70	−110	−110	−110	−115	volts
Zero-Signal D-C Plate Current	200	200	200	200	220	220	ma.
Zero-Signal D-C Screen Current	5	5	9	9	9	9	ma.
Maximum-Signal D-C Plate Current	500	500	800	760	800	780	ma.
Maximum-Signal D-C Screen Current	35	30	48	51	43	41	ma.
Maximum-Signal Peak R-F Grid Voltage	75	70	110	110	110	110	volts
Maximum-Signal Power Input	1000	1000	1600	1900	2000	2340	watts
Maximum-Signal Useful Power Output[3]	610	660	1020	1160	1280	1540	watts

[1] D-C current values shown are for peak conditions, or for single-tone modulation at full signal.
[2] Approximate value; adjust to give stated zero-signal plate current.
[3] Peak envelope power delivered to load from typical amplifier. "Plate power output" is about 10 per cent higher than figures given.

Fig. 6.9B—Typical operation characteristics for the Penta PL-172 operating class AB$_1$.

low grid impedance makes neutralization unnecessary. The grid is supplied with approximately -110 volts of bias, from a low impedance source. The actual voltage is adjusted to produce a zero signal plate current of approximately 220 ma. The screen grid is connected to a source of 500 volts and the suppressor is 75 volts more positive than the cathode. Actually placing this voltage on the suppressor is "gilding the lily" since it will run a cool kilowatt with this element grounded. However, in commercial applications it will produce an increase in power input.

The output circuit is a conventional pi-network but uses a Barker and Williamson tank circuit that is designed to work with high "C" linear amplifiers. Because of the low impedances involved, extra coupling capacitors are used between the tube and the tank circuit. Jennings wide range vacuum variable capacitors are used to cover a variety of loading conditions. An r-f choke across the output jack provides a short circuit path, should one of the pi-network coupling capacitors break-down.

Metering

The grid circuit meter might seem unnecessary or at least a little large for the job. This range was included in the design so that the amplifier could be used for cw applications (class C). A sensitive meter to indicate grid current when operating class AB, would be unnecessary in this rig. If it indicated grid current flow, you would be operating illegally or not loaded heavily enough. The other meters will inform you of these conditions.

Parts List for the deluxe PL-172 pentode linear.

B1—Blower, 117 vac., 30-60 cfm.
C1 C6, C7, C8, C9, C10, C12, C13—.001 mfd., 600 volt disc ceramic cap.
C2, C3, C4, C5—.001 mfd., 600 volt feed-through (Erie)
C14, C15, C16, C17, C18—.001 mfd., 5000 volt (Centralab 852)
C19—5- 500 mmfd. vacuum variable capacitor (Jennings UCSL 500- 3KV)
C20—20- 200 mmfd. vacuum variable capacitor (Jennings UCSL 2000- 2KV)
Cs—Four .001 mfd. (part of PL-184 socket)
J1, J2—Coaxial connector (Amphenol UHF style)
J3—High voltage feed-through connector (Millen 37001)
L1—Low 5-band tank circuit (B&W 852)
LA-1—#47 pilot lamp
RFC-1, RFC-2—2.5 Mh. r.f. choke (J. W. Miller 4537 or equiv.)
RFC-3—Transmitting choke, 500 ma. (B&W 800)
R1—50 ohms, 40 watts. Made from 20- 1K, 2 watt carbon resistors in parallel
S01—Socket for PL-172 (Penta Labs PL-184)
T1—Filament transformer, 6.3 v.a.c. @ 10 amperes (Triad F-21A)
V1—PL-172 tube (Penta Labs)
Miscellaneous components required: Meters—0-20 ma., 0-50 ma., 0-1 ampere (Marion MM2 style), 2—Counter dials (E. F. Johnson Co. 116-208-4), Badswitch knob (E. F. Johnson Co. 116-260), Panel 10½" x 19" (Calif. Chassis PA-15), Chassis 13" x 17" x 1" (Calif. Chassis A-126 modified per text), chassis box 6" x 6" x 6" (L.M.B. UC-973 modified per text) pilot lamp holder (E. F. Johnson Co. 147-206), 8 pin power connector, meter shields (see text), ⅜"-¼" flexible coupling.

The deluxe kilowatt amplifier operates class AB_1 and can run 2800 watts input on a commercial basis, or 2000 watts PEP input in Amateur service.

The plate current meter is connected in the negative power supply lead to avoid the high voltage insulation problems. It should be pointed out that the high voltage bleeder must be connected to the negative high voltage buss, rather than chassis ground. If this precaution is not observed, the plate current meter will also indicate the bleeder current and produce some confusing readings! Proper readings for the PL-172 are shown in fig. 6.9-B.

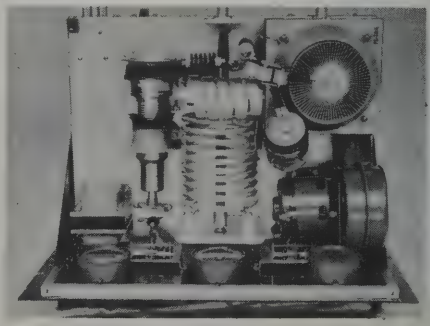
Blower

Although a blower capacity of 50 cfm is required for the 1000 watt dissipation level, 20 cfm appears to be more than adequate at Amateur powers. Some have advocated turning off the air during stand-by periods, but I do not recommend this. I have seen glass seals crack just from the filament heat.

Construction

The major components are mounted on a 13″ x 17″ x 1″ chassis. This method of construction was used to keep all the parts below a 10½″ height limitation. It did necessitate cutting a commercially available chassis down to a 1″ height, however. The PL-172 and PL-184 socket are mounted on an LMB "Utility Box" that originally measured 6″ x 6″ x 6″. Again, this box was cut down to 3⅝″ so that the anode connector of the PL-172 would be near the tank circuit components.

Inside the box, you will find the input circuit components. The r-f power input connector was placed on the rear apron and the 2 watt carbon resistors were grouped around the "hot" terminal. The cathode pins (1, 4, and 7) are connected together and grounded with short leads. An additional ground, made of shielded braid, is connected to the side of the box. All power leads entering the box are effectively bypassed with Erie feedthrough capacitors. Note that two parallel connected feedthroughs are used to handle the heavy filament current. The whole compartment is made air-tight by using Scotch (electrical tape- that is) #33 on the metal edges. Thus, the air from the blower (mounted just forward of the box) can only flow



Top view of the deluxe linear amplifier. Note that flat copper strap is used to wire the antenna connector, the plate cap, coupling capacitors, and the PLATE TUNE vacuum variable capacitor.

Contrary to usual construction, the blower is mounted forward of the tube to place the plate cap near the tank coil. The "plate tuning" vacuum variable capacitor, visible in this photograph, is mounted on a bracket and connected to the dial with a flexible coupling shaft.

through the cooling vanes of the PL-172. A rubber gasket is used at the blower coupling also.

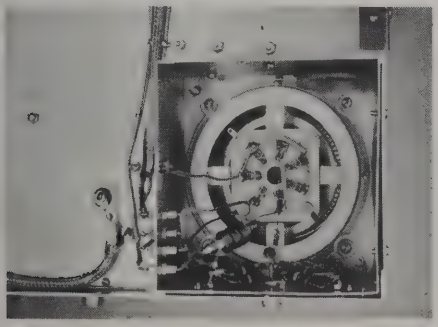
The coupling capacitors (C14-C17) are mounted between two aluminum plates that form a diamond. One end of this combination is connected to the r-f choke, while the other side is mounted on the rear of the tank coil. The input capacitor (C19) is mounted on a bracket, which cannot be seen in the photograph, and connected to the counter dial through a flexible rod. The loading capacitor (C20) is mounted on stand-offs on the left side of the tank, and is connected to the dial through a 3/8"-1/4" flexible coupling. The extra space on the chassis (behind the filament transformer) has been "reserved" for future experiments with the G2MA linear configuration. (See section 6.7).

TVI Precautions

The usual precautions (blocking meters, cage cover, bypasses and shielded wire) were incorporated in the design with the thought in mind that "an ounce of prevention is worth a pound of cure". The extra work may not have been necessary however, since class AB_1 amplifiers do not draw grid current (which "clips" the r-f cycle) or generate TVI producing harmonics, when operated properly. About the only form of interference this amplifier can create is due to overload. No amount of shielding can stop this type of TVI, except shielding the power line, from reaching the power transformer! The shield cover was constructed from Reynolds do-it-yourself aluminum, available in most hardware stores.

Power Supply

A few words regarding the power supply for this amplifier are in order. Any high voltage source between 2000 and 3000 volts will be suitable for the anode voltage. Tests were made on the amplifier with 2600 volts applied. The screen supply should be quite "stiff", that is, have a low impedance output. Electronic regulation is not necessary as long as the class AB_1 mode is faithfully observed. For



The input circuit of the tube is enclosed in an air tight box. Power connections enter the box through small feed-through capacitors. On the rear apron, near the coaxial input connector, will be found the 20 lk, 2 watt carbon resistors that serve as the exciter termination.

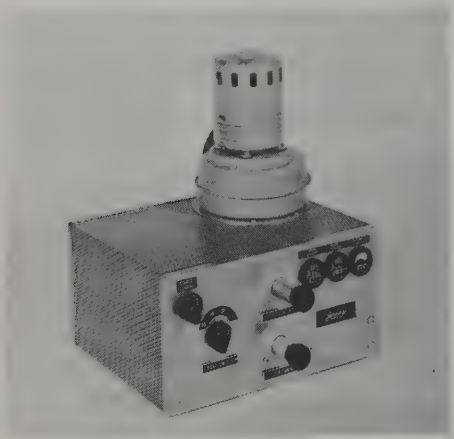
most Amateur applications, the suppressor can be grounded as stated earlier. If you desire to wring the last watt out of the amplifier, a positive 75 volts applied to the suppressor will increase the power output by 10%, or so. The screen requires 500 volts and kicks between 10 and 50 ma necessitating a well regulated choke input supply. The grid bias supply should be adjustable over a range of -40 to -150 volts. A *well filtered* half wave supply, using one of the small 50 ma 1:1 ratio power transformers will be very adequate.

As stated earlier, the bleeder resistor should be "grounded" to the negative high voltage lead and *not* chassis ground. The 117 volt control circuits should be arranged so that the blower comes on with the filaments. A 200 ohm, 50 watt rheostat should be connected in series with the filament transformer line (pin 5) so that the PL-172 voltage can be accurately set to 6.0 volts.

I would like to take this opportunity to thank my father, Lew Stoner-K6BLC for the beautiful construction job he did on this amplifier which, I might add, is characteristic of everything he does.

6.10 THE LITTLE LINEAR

The old adage "good things come in small packages" was never more true, than in the case of the "Little Linear". This 2 kw PEP linear amplifier was designed by Jo Emmett Jennings, W6EI, of Jennings Radio, San Jose, California. This table-top kilowatt AB_1 linear measures approximately 5½" high, 11" long, and 7" deep, and is believed to be one of the smallest amplifiers ever constructed. The main approach to this miniaturization centered around the use of the high-power ceramic Eimac 4CX1000A tube designed especially for class AB_1 operation. This tube is used in conjunction with small Jennings vacuum variable capacitors in the output pi-network and a small Jennings vacuum relay for transmit-receive. Miniaturized meters were also used and they measure plate current, screen current, and antenna current. The band switching is designed for four bands, although by using another switch position, five bands could have been included. The amplifier tunes the 15, 20, 40, and 80 meter bands.



The foundation for the "Little Linear" is the chassis supplied with the California Chassis Co. ham cabinet. For this application, it was inverted and the ridge cut off to make a pan for mounting the components.

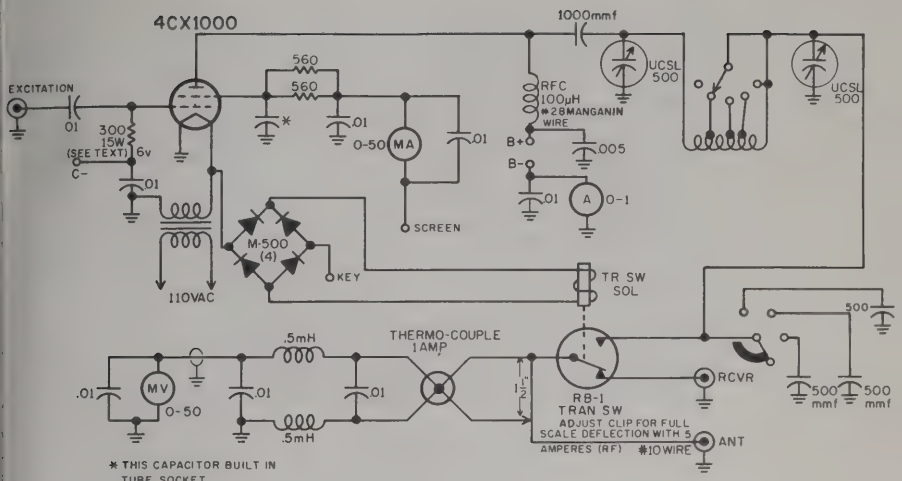


Fig. 6-10A—Diagram for the W6E1 "Little Linear". This amplifier runs 2 kw PEP and is contained (less power supply) in a cabinet *smaller* than the KWM-1.

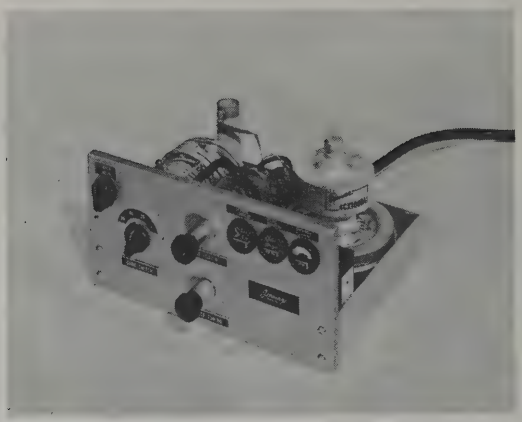
Circuit Details

Two vacuum capacitors form a portion of the pi-network. The 500 mmfd input capacitor is sufficient to tune the desired frequencies. The output capacitor (also 500 mmfd) is used in conjunction with a set of small mica capacitors. The size of the tuning circuitry was reduced by about 16 cubic inches by using this system. In addition to the reduction in size, this setup resulted in considerable savings in vacuum capacitor cost. The UCSL-2000, often used in this service, is considerably larger and more expensive than the UCSL-500's used in this design.

This amplifier also uses the "non-inductive" input system. R-F energy is applied to a 300 ohm, 15 watt, carbon resistor. This resistor may be made somewhat larger or smaller, depending on the driving power available. If the amplifier is used with a "100 watt exciter" a 50 ohm resistor (like section 6.9) should be used.

The output r-f choke for the plate circuit is made with manganin wire, wound on a form approximately one inch in diameter and five inches long. The resistance of this material (#28 d.c.c.) produces a very low "Q" and discourages oscillation which could occur in a resonant choke. This solenoid-type choke has been found to work very well over the frequency range of the amplifier and does not contain "holes" that are characteristic of other chokes.

A thermo-couple is installed in the amplifier to indicate when maximum output occurs. This is necessary, for minimum plate current does not usually correspond to maximum antenna current in tetrode amplifiers. In use, the amplifier is tuned for normal conditions and then re-peaked for maximum antenna current. The extra output is generally worth the effort. The meter does not accurately measure antenna current but simply indicates when the maximum occurs. The schematic (fig. 6.10-A) indicates a movable clip on the antenna lead which acts as a shunt to keep the meter on scale.



The W6EI "Little Linear" uses an Eimac 4CX1000A and runs two kilowatts PEP input. To give you an idea of the size, the case is about one inch smaller, all the way around, than a KWM-1!

The transmit-receive switch is built into the amplifier so that it can be near the pi-network output. Coax connectors to the receiver, exciter, and antenna are all brought out at the rear of the chassis. A simple bridge rectifier circuit, consisting of four Sarkes Tarzian M-500's supplies the necessary dc to operate the coil in the Jennings RB-1 vacuum relay.

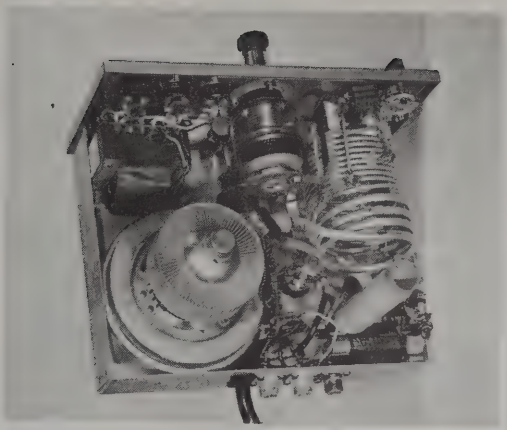
The amplifier is designed to be operated from a remote power supply. A pair of voltage regulator tubes are employed in the power supply to maintain the 4CX1000A screen voltage at a nearly constant potential. Silicon rectifiers are used in the supply in place of the usual vacuum tubes.

The air-cooling used on this amplifier is rather unique in that an exhaust system is employed rather than the conventional blower system. All of the air is not required to pass through the tube socket but does, of course, pass through the anode cooling fins of the 4CX1000A. The entire cabinet and all components are kept much cooler because of the perforated enclosure allows free passage of air from all sides and the bottom. This system has proved itself to be quite efficient since air is not trapped in any area due to turbulence. The cabinet, it should be mentioned, is from the new California Chassis Company, (Lynwood, California) line of perforated ham cabinets.

Tuning

In tuning the amplifier, the grid current is first adjusted to zero and the screen current is maintained as close to zero as possible. As the loading is increased, the plate current starts to rise and the screen current will tend to go negative. The grid drive is then increased until the screen current and grid current are essentially zero. This procedure of gradually increasing drive and increasing coupling to the antenna is carried on until normal plate current is reached.

When operating, the screen current will vary from zero to negative during modulation. Should it start to go positive, a small amount of TVI may be en-

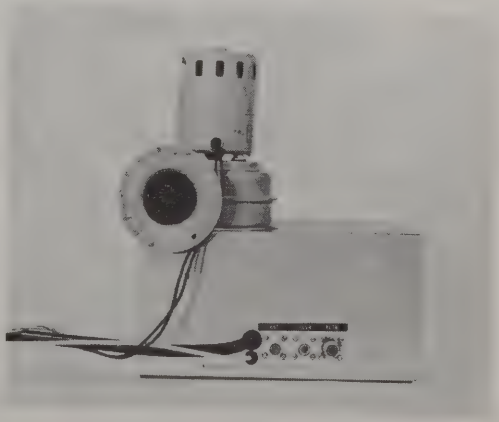


The filament transformer is located just forward of the 4CX1000A, with the meters above. The tank coil, on the right, was made from Air Dux (Illumitronic Engineering) coil stock. Directly behind the tank is the vacuum relay, r-f choke and input swamping resistor. The thermo-couple is partially hidden, to the right of the r-f choke.

countered. If operated as described, however, no TVI problems should occur.

Loading of the grid is accomplished by means of a resistor and capacity coupling to the coax input. This resistor value should be made to correspond to the amount of drive that is available. As an example, if a very high drive level is available, it is necessary to use approximately 50- 100 ohms of swamping. If only a small amount of drive, such as 3- 5 watts, is available, a value between 1000 and 5000 should be used. The resistance across the circuit eliminates the r-f choke which could be a possible source of parasitics. It also stabilizes the amplifier and loads the driving circuit, increasing and maintaining better linearity.

Power and signal circuits are brought out the rear apron. The "blower" is an inhaler, rather than an exhaler.



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7

Receiving Sideband

PROBABLY the biggest single factor holding back single sideband (if it is being held back) is the difficulty in tuning signals. It cannot be denied that AM is easy to receive even when the signal drifts several kcs (which they often do!). The reception of sideband requires entirely different techniques and there is an "art" that must be acquired when using a communications receiver on the sideband frequencies. Undoubtedly there would be many more hams on single sideband if it were as easy to tune as "ancient modulation". But this seems a small price to pay for all the advantages that are gained with single sideband.

The difficulty in tuning signals stems from the fact that the carrier is not transmitted. For proper demodulation, the carrier must be inserted *properly* at the receiver. Properly is the key word, for this is where the problem arises. The reinserted carrier must be very close to the frequency of the suppressed carrier. If the frequency difference is more than 50 cycles or so, the voice will sound unnatural, but the signal will be intelligible until the difference is several hundred cycles. In addition, the amplitude relationship between the sideband and the inserted carrier must also be correct. If there is more sideband energy than carrier (reinserted) the effect will be the same as overmodulation and distortion will result. If the carrier is much stronger than the sideband, it results in under modulation. It has been shown that if the carrier is many times the sideband level the reception will be greatly improved. It is not uncommon for the reinserted carrier to be as much as 100 times the amplitude of the sideband. This insures that the demodulated signal will not be over modulated.

7.1 TUNING IN SSB SIGNALS

Contrary to opinions otherwise, it is possible to copy SSB on *any* Amateur communications receiver. The degree to which they can be copied will depend on how good the receiver is with regards to selectivity and stability. It is possible to demodulate the SSB signal in either of two ways.

Front End Carrier Insertion

The receiver is operated in the usual manner for phone reception; that is avc turned on, bfo turned off, and the r-f gain control full-on. The transmitter vfo a BC-221 frequency meter, or similar auxiliary oscillator is used in conjunction with the receiver in the following manner: Center the SSB signal in the pass-band of the receiver i-f by tuning for maximum upward swing on the "S" meter. Then tune the auxiliary oscillator slowly through the frequency occupied by the SSB signal. You will find that the signal will pass through the characteristic "Mickey Mouse" high pitched sound to the guttural "Mortimer Snerd" sound. In between these two extremes lies the narrow margin of intelligible copy. If by chance you place the reinserted carrier on the wrong side of the SSB signal, the region between the high and low-pitched sounds will not yield understandable output. If this is the case, move the oscillator to the other side of the signal. You may have been trying to copy a lower sideband signal as an upper sideband signal, or vice versa.

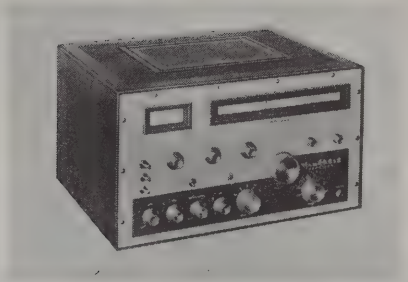
Here is a rule-of-thumb to remember. To tune in a lower sideband signal the carrier must be on the high frequency side, and to tune in an upper sideband signal the carrier must be on the low frequency side of the SSB signal. Assuming that the carrier is placed correctly, it is possible to tune the receiver back and forth across the combination of the SSB and reinserted carrier signal for optimum receiving conditions.

The amplitude of the reinserted carrier should be adjusted by varying the oscillator coupling to the receiver antenna terminal. This will insure that maximum audio is recovered from the SSB signal with no signs of distortion due to over-modulation of the carrier. If heavy interference is encountered, increasing the carrier injection will prevent "capture effect" of the second detector by other strong carriers. Often it will be noted, when using a transmitter vfo for the carrier, that the inserted carrier is too strong and cannot be reduced. But unless the signal is paralyzing the receiver, it will be possible to recover the audio by turning the volume control up beyond the normal point. Incidentally, it might indicate that the TVI suppression of the transmitter is not too good, if that much vfo signal is leaking outside the case.

The crystal filter may be used in a normal manner to notch out heterodynes and to reduce the receiver bandwidth. It is possible to tell which sideband is being transmitted by determining to which side of the carrier the receiver must be tuned in order to derive the maximum audio output from the signal. The receiver must be reasonably selective to do this, however. When using the front end system, the transmitter vfo (or whatever the carrier source is) must have good stability. The receiver stability is relatively unimportant.

BFO Carrier Insertion

When using this method of demodulation, stability of the receiver conversion oscillator(s) and the beat frequency oscillator is a *must*. Here is how you can copy SSB signals using the receiver bfo for the reinserted carrier: With the receiver in the conventional AM position, tune the "S" meter for maximum upward deflection. Turn off the avc, back off the r-f gain, turn up the audio gain to near full on, and lastly turn on the bfo switch. *Slowly* tune the bfo pitch control (*not* the tuning knob) across the receiver pass-band. At one point in the rotation the reinserted carrier should correctly demodulate the SSB signal and the "Donald Duck" will



The Heath "Mohawk" receiver uses 50 kc i-f's to obtain the proper selectivity for SSB reception. In addition, a bridged "T" 50 kc filter is used to null out unwanted interference.

come out as English. Advance the r-f gain until a comfortable level is available. Too much gain will allow more SSB signal to reach the detector than arrives from the bfo, and overmodulation is the net result. It will then be difficult to demodulate the signal no matter where the b.f.o. is placed. The reason for turning off the a.v.c. is that the "shotgun bursts" of sideband cannot be coped with by the time constant of the avc. Thus it is necessary to cut the legs out from under the SSB signal until it fits the receiver.

With this system the tuning rate of the receiver is very important, unless you care to repeat the above procedure each time you change stations. One way to "beat the rap" is to connect a small trimmer that can be adjusted from the front panel. It acts as sort of a vernier or fine tuning.

A Word About Selectivity

A receiver bandwidth of 3 kc is all that is needed to copy *good* AM or SSB A great many AM stations actually have a considerable amount of FM component. When they are received on a broad device, the effect is usually unnoticed. But, when heard on a selective receiver the outcome is pretty un-intelligible. For best results, when copying *clean* AM or SSB signals, the receiver should have no more than a 3 kc bandwidth. In addition the skirt selectivity should be as sharp as possible. If the skirt selectivity is poor, it is possible for an adjacent sideband station to appear to be causing interference (more about this later). With good skirt selectivity it is entirely possible for two sideband stations to operate on the same suppressed carrier frequency without causing interference to each other. Naturally, each one operates on the opposite sideband. In general, comments regarding the selectivity of filter exciters (Filter Chapter) also apply to receiver selectivity. For optimum utilization of spectrum space the receiver and transmitter selectivity should match each other as closely as possible.

Apparent Broadness of Signals

Many AM operators complain of excessive bandwidth of SSB signals. Unfortunately this is occasionally true. But usually the fault does not lie with the SSB signal, but rather, the manner in which the receiver is being used. Most AM operators use their receivers with the avc turned on and the r-f gain wide open. When they are listening to a station in a "channel" next to a SSB station, or not copying any station, the avc voltage is nil and the receiver gain is maximum. The non-existence of avc voltage raises the gain of the r-f and i-f stages so that the *effective* bandwidth of the receiver is perhaps 15 or 20 kc, rather than the 5 or 6 kc measured by the manufacturer at the 3 db points. Naturally, when the adjacent channel SSB signal starts up, it will be heard and quite loudly! The "jumping-up-and-down" of the SSB signal creates a terrifying effect on the avc system (not to mention the AM operator). If the SSB signal had been an unmodulated carrier of the same peak amplitude, the receiver avc voltage would have been raised and held there. This would result in reduced receiver gain and a

decrease in the effective bandwidth. Interference from an adjacent SSB station can be reduced considerably, simply by turning off the avc voltage and reducing the r-f gain control well below the point of overload. Try it—you may be surprised to find that the SSB signal is not so broad after all!

7.2 ADAPTERS FOR RECEIVING SSB

There are three prime requirements that must be present in any *good* single sideband receiving system: (1) Stability (2) Selectivity—a 3 kc bandwidth with rejection outside this range (good skirt selectivity), and (3) Exalted carrier reception. This means having the bfo voltage at least 10 to 100 times as great as any other signal that is present in the second detector. This reduces the "monkey chatter" caused by strong signals cross modulating with each other upon detection.

There is no need to dwell on the first requirement. Either you have a stable receiver, or you do not. Often this malady is unbeatable and the owner winds up buying a new communications receiver. Requirement two can be incorporated in older receivers with excellent results. The most desirable place to achieve selectivity in a receiver is near the first mixer, before the i-f amplifier. Pre-intermediate frequency selectivity is desirable because it prevents strong adjacent channel signals from producing distortion products in the latter stages. These distortion products appear as splatter on the signal but are actually transmitted by the "offending" stations. There are several means of obtaining pre- i-f selectivity such as crystal filters, Burnell toroid filters, and Collins Mechanical filters.

Post i-f selectivity is becoming quite popular. The "Lazy Man's Q-5er" and the "SSB Q-5er" are the most popular of the double conversion attachments. The heart of these adapters is the surplus BC-453 Command Set receiver. Although the skirt selectivity is not as good as the filters mentioned earlier, a 6 db bandwidth of 2.7 kc makes the BC-453 very effective for demodulating sideband signals. The bfo in the outboard receiver furnishes the reinserted carrier and is placed on either side of the 85 kc pass-band, depending on which sideband is being received. The BC-453 has no avc buss and therefore the r-f gain control must be used to control the signal level, thus avoiding overmodulation.

Another post i-f adapter is known as the slicer. In this unit, i-f signals are supplied from a source just preceding the detector. The slicer looks like a phasing exciter "turned inside out". By proper phasing, one sideband is rejected while the other sideband is amplified. The Central Electronics Model A and B Sideband Slicer (trademark) and the Lakeshore Signal Splitter are such units.

Parts List

R1—10,000 ohms, ½w.
R2—1000 ohms, 1w.
R3, R4, R22, R23, R24, R25—3,300 ohms, ½w., 5%
R5—10,000 ohms, ½w.
R6, R7—1 megohm, ½w.
R8, R10—51,000 ohms, ½w., 5%
R9—22,000 ohms, ½w.
R11, R12—3.3 megohms, ½w.
R13—680,000 ohms, ½w.
R14, R15—2,000 ohms, ½w., 1% precision
R16, R17—7,000 ohms, ½w., 1% precision
R18, R21—133,300 ohms, ½w., 1% precision
R19, R20—100,000 ohms, ½w., 1% precision
R26, R28, R29, R31—220,000 ohms, ½w.
R27, R30—100,000 ohm pot.
R32—100,000 ohms, ½w.
R33—5,600 ohm, ½w.
R34—47,000 ohms, ½w.
R35—2.2 megohms, ½w.
C1—140 μμfd. variable
C2—250 μμfd. mica
C3—0.008 μfd. mica
C4—0.003 μfd. mica
C5—0.001 μfd. mica
C6—0.05 μfd. paper
C7—100 μμfd., 5% mica
C8—75 μμfd., 5% mica
C9—50 μμfd. variable
C10, C11—110 μμfd. mica (matched within 1%)
C12—470 μμfd. mica
C13, C14—0.01 μfd., mica or paper
C15—2,340 μμfd. (paralleled 0.002 μfd. and 170-780 μμfd. trimmer)
C16—4,860 μμfd. (paralleled 0.0043 μfd. and 170-680 μμfd. trimmer)
C17—607.5 μμfd. (paralleled 500 μμfd. and 9-180 μμfd. trimmer)
C18—1,215 μμfd. (paralleled 0.001 μfd. and 50-380 μμfd. trimmer)
C19—0.1 μfd. paper
C20—500 μμfd. mica
C21—0.25 μfd. paper
C22—0.1 μfd. paper
C23—10 μμfd. mica
J1—open circuit phone jack
L1, L2, L3, L4—National R-100 r-f choke modified so that second serves as secondary (L2). The remaining pi's are connected in series as shown.
PS-1—two channel 90° phase shift network.
S1a, S1b—2-pole, 4-position wafer switch.

The RME Electro-Voice Model 4301 adapter looks more like an ordinary speaker cabinet, but contains the "tail end" of a single sideband receiver.

7.3 A PHASING ADAPTER

The phasing type of adapter was originally described by Norgaard (GE Ham News—July Aug. 1951) and subsequently a more commercial version designed by Wes, W9DYV, was introduced as the Central Electronics Model A "Slicer". The basic circuit for the slicer is shown in fig. 7 3-A, and is complete except for the power source. This unit uses the phasing principle to discriminate between sidebands of a received signal. It should be noted that it is equally useful for eliminating heterodyne interference on AM as well as SSB stations.

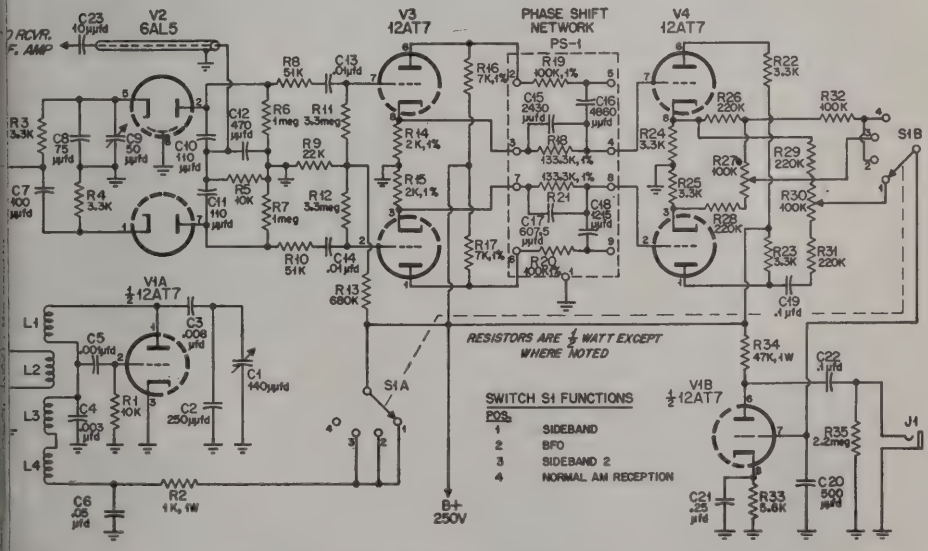


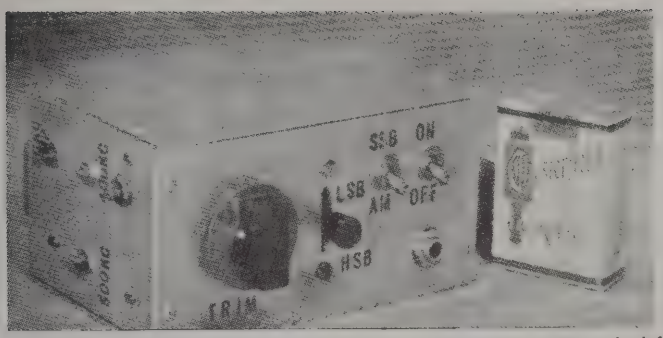
Fig. 7.3A—Phasing type receiving adapter. By using suitable r-f and a-f phase shift networks, the upper sidebands can be made addative, while the lower sidebands cancel out. By switching the "adder" (V_4) phase the opposite condition will prevail, thus providing sideband selection.

The Central Electronics "Sideband Slicer" circuit is shown in fig. 7.3-A. This "B" Model, however, includes a "Q" Multiplier for interference rejection.

The principle of this slicer is very similar to that of the phasing type single sideband exciter. The incoming i-f signals are detected by the two diodes in the 6AL5 by mixing with the r-f voltage from the 455 kc oscillator. This r-f voltage is divided into two equal parts and shifted in-phase with respect to each other by 90° in the R/C networks (R3-C8, C9 and R4-C7). The two separate signals (after detection) are passed through the phase-shift network (PS-1) and then combined in the output of V4. One stage of audio amplification (V1b) is included. Depending on the position of the switch (S1), the signals on one side of the 455 kc oscillator frequency will be combined in such a way as to be cancelled out, while the signals on the other side of the oscillator will be reinforced and heard in the headphones. This unit effectively splits the receiver selectivity characteristic in half, rejecting the one half while receiving the other.

The operating principal of the Lakeshore "Signal Splitter" is the same as that described in section 7.3. This adapter features extensive use of printed circuit boards.

The Technical Materiel Corporation GSB-1 employs a toroid filter operating at an i-f of 17 kc. A self contained audio amplifier eliminates the need for extensive receiver modifications. An interesting avc circuit, used in this adapter, is discussed in section 7.6.



Variable mica padders for 50 kc amp. and 400- 500 kc osc. tuning are on the left side while "Trim" control varies 400 or 500 kc osc. for fine tuning of SSB signals.

If the reader desires to duplicate this device, the original article should be consulted for detailed construction information and for the alignment data. The complete phase-shift network, PS-1, is available from Central Electronics and the James Millen Co. The complete adapter is available from Central Electronics wired or in kit form.

7.4 THE TRANSLICER

The "Translicer" is another form of post i-f sideband adapter. The name is derived from the fact that this device uses no vacuum tube, but instead uses low cost- easily obtained transistors. Transistorizing the circuits results in a unit of much smaller size and lower power drain than could be accomplished with vacuum tubes.

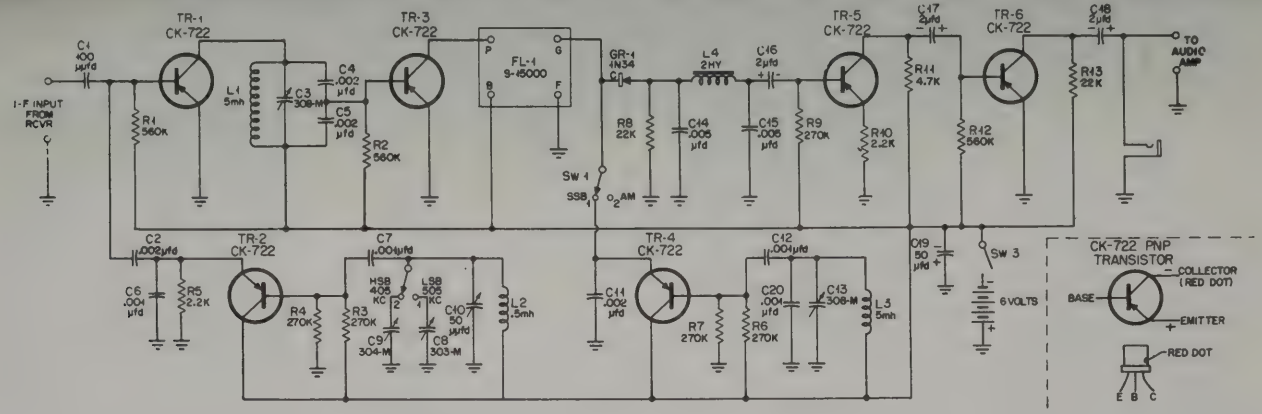
The highly selective element is a 47-50 kc sharp sideband filter, Burnell type S-15000. The carrier side of this filter has an attenuation rate comparable with a 250 kc mechanical filter, while the opposite slope has a somewhat lower attenuation rate which results in a smoother high frequency roll-off and hence a more agreeable voice quality than a sharply breaking filter. It should be realized that when switching sidebands in the circuit described, the inversion is such that the steep side of the filter is always doing the work of rejecting the opposite sideband.

Circuit Description

Functionally the circuit (fig. 7.4A) may be described as follows. A CK-722 transistor input stage, TR1, acts as a frequency converter, mixing the 455 kc i-f signals from the radio receiver with either a 405 kc or 505 kc signal from the local oscillator, TR2 (CK-722). Switching the oscillator frequency between 505 kc and 405 kc produces lower and higher sideband outputs, respectively. The 50 kc i-f produced in either case is amplified by TR-3 before being applied to the sideband filter, FL-1. The output of the filter is combined with a 50 kc carrier frequency oscillator and demodulated to audio by the diode, GR-1. The 50 kc oscillator circuit is rather unusual in that the emitter return is through the output coil of the filter which has a low dc resistance but high reactance at 50 kc. This results in very effective mixing of the carrier with the filter output signals.

Following the diode is a 3000 cycle low pass filter of conventional design. Two audio stages follow the LPF. These are both grounded emitter stages. The resulting audio output is of comfortable earphone level. In most cases the output will be applied to the audio amplifier of the associated receiver for loudspeaker operation.

AM operation, on either sideband, is obtained by opening the emitter return of the 50 kc carrier oscillator, TR4.



Parts List

C1 — .0001 µfd., 500V., Silver Mica, 5% (CM15-E-101J)

C2, C4, C5, C11 — 002 µfd., 500V., Silver Mica, 5% CM20-202)

C3, C13—425 To 1260 µµfd., Variable Mica (El Menco 308-M)

C6, C7, C12, C20—.001 µfd., 500V., Silver Mica, 5% (CM20-102)

C8 — 65 To 320 µµfd., Variable Mica (El Menco 303-M)

C9—100 To 500 µµfd., Variable Mica (El Menco 304-M)

C10—50 µµfd., Variable Air (Hammarlund HF-50)

C14, C15 — .005 µfd., 500V., Mica, 10% (CM30-502)

C16, C17, C18 — 2 µfd., 12V., Electrolytic (Barco PT12-2)

C19—50 µfd., 25V., Electrolytic (Barco P25-50)

R1, R2, R12 — 560,000 Ohms, ½ W., Comp., 10% ("Little Devil")

R3, R4, R6, R7, R9 — 270,000 Ohms, ½ W., Comp., 10%

R8, R13 — 22,000 Ohms, ½ W., Comp., 10% ("Little Devil")

R5, R10 — 2,200 Ohms, ½ W., Comp., 10% ("Little Devil")

R11—4,700 Ohms, ½ W., Comp., 10% ("Little Devil")

L1 — Inductor, Choke, 5 mh. (Miller #954)

L2 — Inductor, Choke, 0.5 mh. (Miller #951)

L3 — Inductor, Choke, 5 mh. (Miller #954)

L4 — Inductor, 2.2 hys., Mean (UTC VIC-13 [or equivalent])

F1—1—47-50 kc., Sideband Filter (Burnell S-15000)

S1, S2, S3 — S.P.D.T., Rotary, Wafer (Centralab #1460)

J1—Jack, Single Contact (Mallory SC-1A)

GR1—Diode, Germanium-1N34

TR1, TR2, TR3, TR4, TR5, TR6—Transistor, PNP (Raytheon Ck-722) Junction

Battery—6 Volts (4—1½ V. Cells) (Eveready #7)

Fig. 7.4A—The "Translicer" schematic diagram. The 455 kc i-f signal is converted to 50 kc and passed through a 3 kc toroid filter. Interference in the region of the unwanted sideband is thus removed.

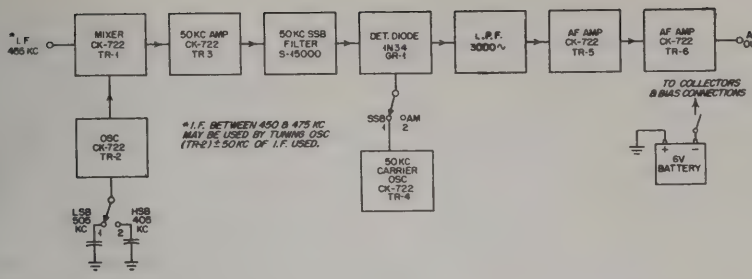

*I.F. 455 KC
MIXER CK-722 TR-1
50KC AMP CK-722 TR 3
50KC SSB FILTER S-45000
DET. DIODE 1N34 GR-1
L.P.F. 3000~
AF AMP CK-722 TR-5
AF AMP CK-722 TR-6
AUDIO OUTPUT
TO COLLECTORS & BIAS CONNECTIONS
6V BATTERY
OSC CK-722 TR-2
*I.F. BETWEEN 450 & 475 KC MAY BE USED BY TUNING OSC (TR-2) 50 KC OF I.F. USED.
SSB
CAM
50 KC CARRIER OSC CK-722 TR-4
LSB 505 KC
HSB 405 KC

Fig. 7.4B—Block diagram of the "Translicer". Note that the transistor oscillator stage operates either 50 kc above or 50 kc below the i-f.

The unit was designed for 6 volt operation since this is convenient from commonly available cells and yields the required output. A self-contained battery of four 1½ volt flashlight cells will give many hundreds of hours of operation and will result in excellent stability.

Since the CK-722 is a PNP junction transistor, the positive side of the battery is grounded and the negative side is connected to the collectors and base biases. For NPN junctions, the polarity would be reversed.

It was found that this particular type of transistor ran quite uniform in characteristics, from unit to unit, and gave good performance when used as a mixer, oscillator and amplifier.

Construction

The slicer, complete with 6 volt battery, was built in a small aluminum box measuring 6" long, 4" deep and 3" high. The 50 kc sideband filter is mounted on the back and extends an additional 2⅛" to the rear.

The input mixer and 400- 500 kc oscillator circuits are mounted on the left side of the box, while the 50 kc oscillator, LPF, and audio stages are mounted on the right. Due to the low levels employed and the lack of ac hum problems, the layout of components is considerably less critical than equivalent vacuum tube circuits. No internal shielding is required. On the front panel from left to right are the 400- 500 kc oscillator trimmer, the HSB-LSB switch, the AM-SSB switch, and the battery switch with a monitoring jack underneath. The variable mica main capacitors for the 400- 500 kc oscillator may be tuned from the outside of the box, as can be seen from the photographs.

The Miller r-f chokes specified, make suitable low cost inductors for the 400-500 kc and 50 kc oscillators. Some improvement in frequency stability would be obtained through the use of the more expensive miniature toroids. The stability obtained with the present coils, however, seems adequate at room temperature, plus or minus a few degrees.

All wiring is straight point-to-point using solid, tinned #20 wire with spaghetti tubing.

When using transistors, it is a good idea to always wire up the rest of the components first and to put the transistors in last. This will prevent undue heating of the transistors while wiring other components. Also, the normally supplied lead lengths should be retained on the transistors. Place small lengths of tubing over each one and fold over as required, but do not cut them off short. You may want to use them over again!

Adjustment and Operation

The carrier oscillator must be adjusted to exactly 50 kc by means of trimmer C13. This may be done by comparing the second harmonic of the 50 kc oscillator

with a 100 kc crystal standard, common in many stations. Switch S1 must be in the SSB position for this, and coupling to the oscillator should be as light as possible.

Next, the i-f oscillator should be adjusted to plus 50 kc of the receiver i-f with S2 in the LSB position and to minus 50 kc of the receiver i-f with S2 in the HSB positions. Trimmers C8 and C9 are used for the LSB and HSB respectively. A signal generator at the receiver i-f is helpful for this, but not absolutely necessary. Once the slicer is operative, headphones may be plugged into the monitor jack and final trimming adjustments may be made on actual signals. An AM broadcast station is very good for final line up of the i-f oscillators. With the switch, S1, in the SSB position, the associated receiver is tuned to zero beat with the AM carrier, as monitored in the slicer headphone jack. Now, switch S2 to the opposite sideband and tune the trimmer for this position to again produce zero beat. When exact zero beat is obtained on both LSB and HSB positions, then the i-f oscillator is positioned exactly correct.

The front panel trimmer is useful for tuning in signals precisely, in lieu of the main tuning control on the receiver. Remember that when using an adapter of this sort the receiver bfo must always be off for any type of reception.

7.5 THE 1959 SSB Q5'ER

The author has always considered the BC-453 Command Set receiver to be an excellent foundation for a SSB receiver. This surplus "goodie" uses a two stage 85 kc i-f amplifier which has a bandwidth of 2.7 kc at 6 db down and produces good results on sideband. The tuning range of the BC-453, however, is 200- 500 kc, which necessitates the use of converters. An article describing this system appeared in the September 1956 issue of CQ Magazine.

In this system, a crystal controlled converter was placed ahead of the BC-453 and reception of the 75, 40, and 20 meter bands was possible. The crystal frequencies were selected to place the converter output within the tuning range of the BC-453. The reception of sideband stations on 75 and 40 was excellent, but because of the low variable i-f, images were very strong on the 20 meter band. Since that time, the design has evolved (and is still evolving) into the 1959 version of THE SSB Q5'ER. The new design tunes five Amateur bands, 80 through 10 meters, and images are almost non-existent. The converter section uses only one crystal and yet is double conversion! In addition, the Command Set receiver has been modified to include a product detector. (See section 7.6.)

The heart of the converter, shown in fig. 7.5-A, is an ingenious oscillator multiplier system. A single crystal oscillates continuously on 3.5 mc for the second

Parts List

C1, C3—7-45 mmfd. rotary ceramic trimmer (Centralab 822BN)
C2, C8, C21—33 mmfd. npo disc or mica
C4, C5, C6, C7, C11, C12, C14, C16, C18, C19—.005 mfd., 600 volt disc ceramic
C9—20 mmfd. npo disc or mica
C10, C13, C15—100 mmfd. npo disc or mica
C20—4-25 mmfd. rotary ceramic trimmer (Centralab 822 AZ)
C22—.001 mfd. disc ceramic
J1, J2—RCA type phono connector
L1—100 turns, #34 enam. scramble wound on ¼" slug tuned form, link 8 turns #34 enam.
L2—15 turns, #26 enam. closewound on ¼" slug tuned form, link 3 turns #26 enam.
L3—10 turns, #26 enam. closewound on ¼" slug tuned form, link 2 turns #26 enam.
L4, L5—15 turns, #26 enam. closewound on ¼" slug tuned form.
L6—10 turns, #26 enam. closewound on ¼" slug tuned form
L7—14 turns, #34 enam. closewound on 1 megohm, 1 watt carbon resistor
L8—24 turns, #34 enam. closewound on ¼" slug tuned form (Miller #4504)
Note: Coils L1 through L6 wound on Miller #4500
R1—220 ohm, ½ watt
R2, R6—100K, ½ watt
R3—4.7K, ½ watt
R4, R7—1K, ½ watt
R5—1 meg., ½ watt
R8, R9—10K, ½ watt
R10—47K, ½ watt
R11—22K, 2 watt
R12—40 ohms, 1 watt
R13—15K, 1 watt
RFC—2.5 mh. r.f. choke (Miller #6302)
S1—6 pole, 5 position ceramic wafer switch, 2 decks, 3 poles per section. (Centralab PA-2021)
T1—4.5 mc. interstage i.f. transformer (Miller #6203)
XTAL—3500 kc. crystal (should be as active as possible)
Miscellaneous components: 6CB6 tube, 6AN8 tube, 12BA7 tube, 7 pin socket and shield, two—9 pin sockets, crystal socket, chassis box

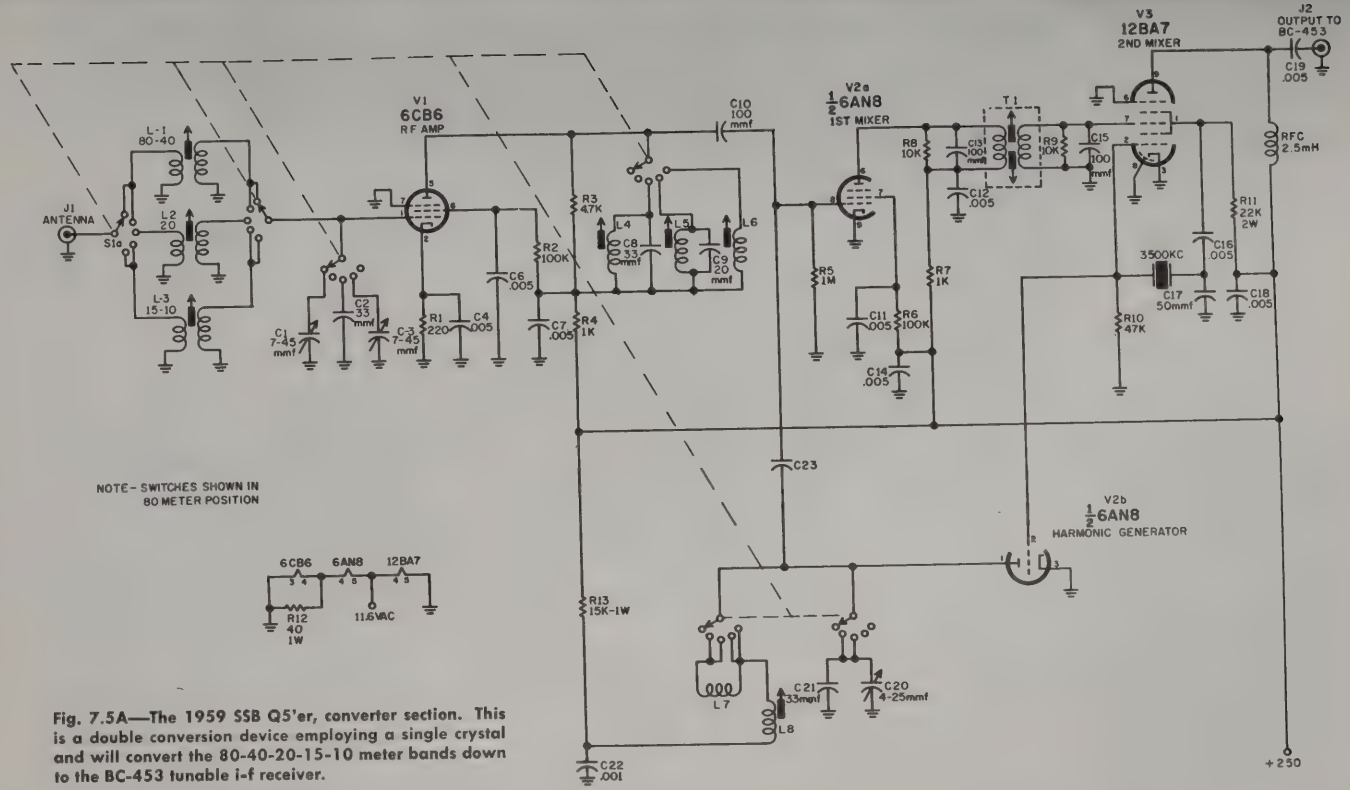
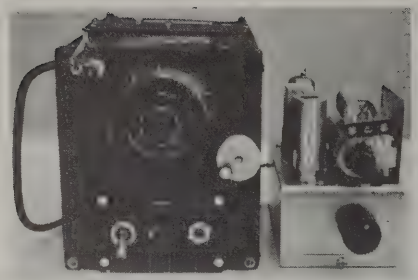


Fig. 7.5A—The 1959 SSB Q5'er, converter section. This is a double conversion device employing a single crystal and will convert the 80-40-20-15-10 meter bands down to the BC-453 tunable i-f receiver.



At the moment, you're face-to-face with the 1959 SSB Q5'er. The converter uses double conversion with both stages crystal controlled and tunes the 75, 40, 20, 15, and 10 meter bands with no images. When used with the BC-453 (which has been modified to incorporate a product detector) it makes a triple conversion communications receiver.

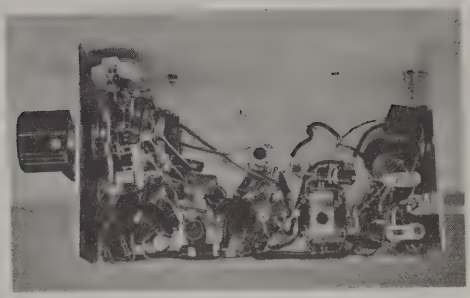
mixer and a harmonic generator produces the correct oscillator frequencies for the first mixer. Rather than start at the beginning, as is often done, let's start at the end—the output to the BC-453. Signals between 190 kc and 550 kc are the only frequencies that can be received by the Command Set. A type 12BA7 pentagrid mixer feeds these signals to the BC-453. The 3500 kc crystal is connected between the screen and oscillator grid and provides injection voltage. The signal grid (pin 7) of V3 is connected to a broadband transformer that accepts signals between 3.7 and 4.0 mc. Thus, if a 3.7 mc signal was applied to pin 7 of V3, it would heterodyne with the 3.5 mc injection signal and produce a 200 kc beat difference. Similarly, a 4.0 mc signal would beat with the oscillator and produce a 500 kc output. Thus, the 75 meter phone band is tuned between 300 and 500 kc on the BC-453. So far this is quite similar to the original SSB Q5'er and the earlier Novice Q5'er. Now, let's jump over to the antenna jack (J1) and "follow" a 75 meter signal through the converter. Note that the switches are shown in the 80 meter position. Phone signals are applied to L1 and are subsequently coupled to the grid of V1, a type 6CB6. In this position of the switch, C1 resonates L1 at 80 meters. The plate circuit of V1 is untuned and the signal is coupled to V2 through capacitor C10. The highly amplified signal passes through T1 and is heterodyned as described earlier. Note that the harmonic generator is disconnected on 80 meters.

Now, let's repeat the process to see how the converter works on 40 meters. Such a signal is still applied to L1 but C1 is now disconnected and the slug in L1 resonates the coil on 40 meters. The r-f amplifier still provides ample gain and the plate circuit is untuned. In the 40 meter position, the harmonic generator is energized, coils L7 and L8 are series connected and resonated by C20 to 10.5 mc (the third harmonic of the crystal frequency). The incoming 40 meter signal (7.2 mc.) beats with the 10.5 mc harmonic in the mixer stage (V2b). The difference signal (3.3 mc) is mixed with the 3.5 mc crystal signal in the 2nd mixer and an output of 200 kc is obtained. Thus, through two conversions, 7.2 mc equals 200 kc.

On 20 meters, the harmonic multiplier remains on 10.5 mc but L2 and L4 are switched into the r-f stage. A signal at, say, 14.2 mc beats with the 10.5 mc energy and produces a V2a output of 3.7 mc. This again beats with the 3.5 mc crystal signal and provides a 200 kc signal for the BC-453. Thus, 14.2 mc equals 200 kc after two conversions. Naturally, 14.3 mc would appear at 300 kc on the dial.

For 15 and 10 meter reception, the 7th harmonic of the crystal frequency is used for oscillator injection at 24.5 mc. L7, C21, and C20 are removed from the circuit to raise the frequency from 10.5 mc to 24.5 mc. Coils L3 (resonated by C3) and L5 (resonated by C9) are used to tune the r-f amplifier. A signal at 21.2 mc would be amplified and mixed with the 24.5 mc energy to produce a beat at 3.3 mc. This, of course, mixes with the oscillator in the 12BA7 stage providing a 200 kc signal for the tunable receiver.

It was stated earlier that the 1959 SSB Q5'ER is still evolving. You will note, if you calculate the frequencies involved in the 10 meter conversion process, that



This should be titled "rats nest" but it's actually the underside of the converter for the 1959 Q5'er. An extra section was added to the bandswitch for the grid coils after this photo was taken.

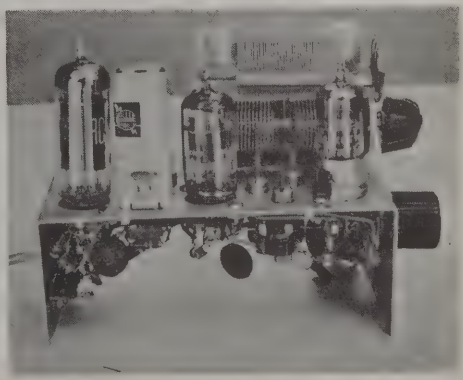
the converter allows tuning only a small portion of the 10 meter band. The author is experimenting with other crystal frequencies to include more of the 10 meter band. With any crystal it will only be possible to tune 350 kc of any ham band because of the receiver range limitation. It should be pointed out that the converter will work with any low frequency receiver. There are certain units of the Navy RA series that would be ideal for a variable i-f receiver.

Another improvement over the earlier SSB Q5'ER is the inclusion of the product detector. Since this is a SSB receiver, the original AM detector was replaced and the bfo switch eliminated. The 12SR7 oscillator/detector was replaced with a 12SN7 and an additional 12AU7 was mounted on the rear chassis ledge. This, of course precludes the possibility of mounting a power supply back there.

Construction

The power supply for the Q5'er was described in the January 1956 issue of CQ. (The Novice Q5'ER). Any supply capable of delivering 250 volts at 60 ma and 11.6 volts ac at 2 amperes will be satisfactory. A Stancor PC-8403 or Triad R-6A is ideal for the job. To obtain the necessary 11.6 volts, the 5 and 6.3 volt windings are connected in *series aiding* and a 6X4 used as a rectifier. The original B plus choke, in the Command Set, is quite adequate to filter the high voltage.

The product detector modifications start by removing the 12SR7 tube and removing all connections from the socket, except 7 and 8 which are filament and ground. The filament pins are the same as the 12SN7. Next, wire the 1, 2, and 3 section of the 12SN7 as the bfo and the 4, 5, and 6 section of the tube to the last i-f transformer. The 12AU7 was mounted in the hole vacated by the dynamotor shock mount, directly behind the 12A6 audio output tube. The sideband switch, S1, must be mounted near the bfo can since a long lead cannot run up to the front



High frequency converter for the 1959 SSB Q5'er. The 6CB6 r-f amplifier is at the right, then the interstage coils, and the 6AN8 mixer/harmonic generator. The "K-Tran" is a 4.5 mc interstage unit, modified to cover 3.5- 4.0 mc. At the rear of the chassis is the 6BA7 second conversion stage. The converter output is 190- 550 kc. The multiband tuner in the background was replaced with individual grid coils to eliminate images on 20 meters. The coils are mounted above the chassis in the spot occupied by the variable capacitor.

panel. A shielded wire would probably have too much capacity. The volume control was also mounted on the side apron near the switch, and all volume adjustments made with the front panel r-f gain control.

The converter originally used a "multiband tuner" in the grid circuit of the r-f stage. However, there were several ghost signals on the 20 meter band due to the two resonances always found in multi-band tuners. The original design was changed to include another switch "deck" and three additional coils. The coils L1, L2, and L3 are mounted on the top-side of the chassis in a horizontal position, and the leads brought down through the chassis to the front switch deck. Any attempt to mount these grid coils under the chassis will result in violent oscillation of the r-f amplifier. The new coils occupy the multiband tuner position, that is shown in the accompanying photographs. The components associated with V1 are grouped tightly around the socket. Decoupling resistor R4, is mounted just forward of the socket. Coils L4, L5, and L6 are mounted between V1 and V2. Harmonic multiplier coil, L8, is seen setting by itself in the photograph. The series coil, L7, is mounted by its own leads on the back of the bandchange switch. The interstage transformer, T1, is mounted between V2 and V3, and the crystal was positioned alongside, in the area of minimum heat. Output jack J2 is mounted on the rear apron. The LMB #136 chassis box provides adequate room with some crowding of parts. A spare parts locker could be mounted in the "extra" room at the rear of the chassis.

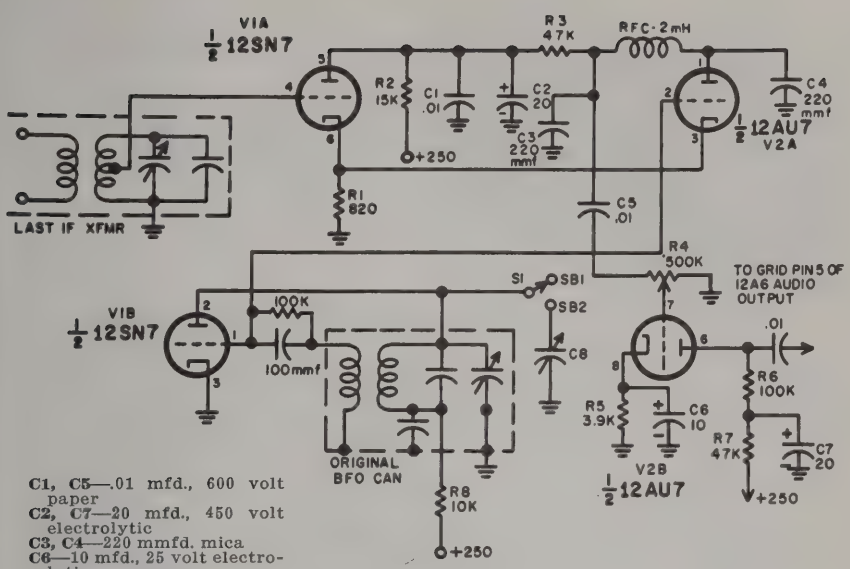
Alignment

When the wiring is completed, remove V1 and V2, and couple an antenna wire to pin 6 of V2 through a 5 mmf capacitor. This will feed signals to the last mixer and broadband transformer. Check for proper oscillation by measuring the voltage at pin 2 of V3. It should be about 30 volts, negative. Tune the BC-453 to 200 kc and peak one slug in T1 for maximum noise or signal. Then tune up to 400 kc and peak the other slug in a similar manner. It may be necessary to fiddle with the slugs to get the frequencies "in the right county" initially. Signals in the 75 meter phone and 80 meter Novice band should "pour in".

Using a grid dipper, set L1 at 7.2 mc, L2 and L4 at 14.2 mc, L5 at 21.3 mc, and L3 and L6 at 28.5 mc *with the bandswitch on the appropriate band.* With the switch in the 10 meter position, set coil L8 to 24.5 mc. Then energize the oscillator and peak L8 for maximum 7th harmonic output. The output will be quite low. For example, it reads only .1 on a Millen grid dipper. Measure the negative voltage on pin 8 of V2. It should be approximately .7 volts with the crystal removed and 1.5 volts with the crystal inserted. At this point, fix the slug in position with wax or gum for it will not require further adjustment. It should now be possible to hear a signal near 28.5 mc. Peak L3 and L6 for maximum signal strength. Next, switch to 15 meters. Weak signals should be heard, and peaking C3 and L5 should bring them in.

On 20 and 40 meters, L7, C20, and C21 are switched into the circuit to lower the harmonic output to 10.5 mc. However, a very strange thing can happen with improper adjustment, and the constructor should be warned of this condition. C20 resonates the L7, L8 series combination. At one point, near resonance, the phase conditions in the harmonic multiplier will be such that the crystal will take off on its third overtone, *in addition to its fundamental oscillation frequency!* This will cause audible whistles to appear at several spots across the dial. Therefore, *do not* adjust C20 for maximum r-f but detune it slightly so this condition does not exist. The amplifier will be perfectly stable and the voltage at pin 8 of V2a will be around minus 2 volts.

Once you are satisfied that the harmonic amplifier is working properly, tune in a 20 meter signal and peak the slug in L2 and L4 for maximum signal strength. Now switch to 40 meters and peak the slug in L1 for maximum signal strength.



C1, C5—.01 mfd., 600 volt paper
C2, C7—20 mfd., 450 volt electrolytic
C3, C4—220 mmfd. mica
C6—10 mfd., 25 volt electrolytic
C8—7 - 45 mmfd. rotary ceramic trimmer (Centralab 822-BN)
R1—820 ohms, 1 watt
R2—15K, 1 watt
R3—47K, 1 watt
R4—500K potentiometer, audio taper (Centralab Radiohm B-60
R5—3.9K, ½ watt
R6—100K, ½ watt
R7—47K, ½ watt
RFC—2 mh. r.f. choke (Miller #6302).
S1—SPST toggle switch (sideband selector)
Miscellaneous components: 12SN7 tube, 12AU7 tube, 9 pin miniature tube socket

Fig. 7.5B—The product detector section of the 1959 SSB Q5'er. A sideband selector switch has been added to eliminate the need for returning the bfo coil.

Note: Original bfo couplings capacitor should be removed.

Lastly, switch to 80 meters and adjust C1 for maximum signal strength somewhere near the center of the band. If cross modulation is noted (broadcast station "birdies") on 80 and/or 40 meters, insert a 220 ohm resistor, bypassed with a .005 mfd. disc, in series with pin 9 of V2a. This will increase the bias on this stage slightly.

7.6 INTERESTING COMMERCIAL CIRCUITS

The commercial receiver manufacturers are very aware of the need for better sideband receivers. Many of them have developed clever circuits that are of interest to experimenters. The author has selected several from the instruction books and they are presented here for your information. No particular construction data is provided, only the basic data for experimentation.

Selectivity

Unless i-f filters or other devices are used, the coupling and number of tuned

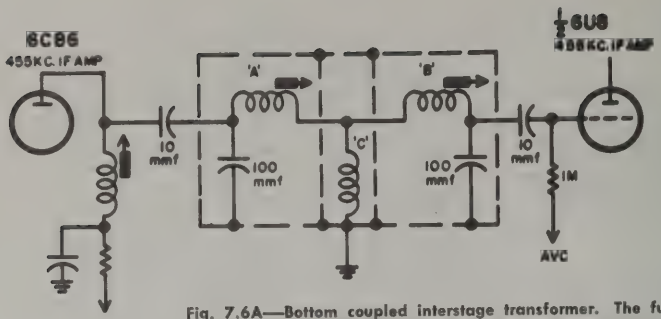
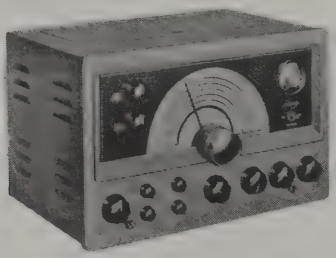


Fig. 7.6A—Bottom coupled interstage transformer. The function of coils "A", "B" and "C" is described in the text.

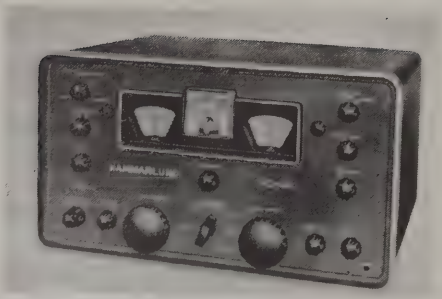
circuits in the i-f amplifier will determine the receiver selectivity. A coupling system, that has been used in high quality video i-f strips for years, is finding its way into Amateur communications receivers. Fig. 7.6-A shows the interstage section of a "bottom coupled" i-f transformer. Coils "A" and "B" are electromagnetically decoupled by shielding or other means. Coil "C" controls the amount of coupling. If coil "C" were a short circuit there would be no coupling and if it were a large inductance the coupling would be excessive. With a small amount of inductance (usually less than 100 uh) critical coupling can be reached. It can also be seen that if such a coupling method was used between each i-f stage, and the coupling coil slugs adjustable from the front panel, it would be possible to construct a continuously variable selectivity i-f system. This "bottom coupled" system is becoming quite popular where variable selectivity is required, and can be found in such receivers as the RME 4350, Hammarlund HQ-170, National NC-303, and Heath "Mohawk".



Bottom coupled i-f transformers are used as interstage coupling devices in the RME 4350 receiver. This coupling method can be used to good advantage in home-brew equipment, as explained in the text.

The "Q" Multiplier is an excellent and simple means of obtaining more selectivity. The Hammarlund HQ-160 employs a "Q" Multiplier, such as that shown in fig. 7.6-B, for this purpose. With such a device, the receiver 6 db bandwidth may be continuously variable from 3 kc to 100 cycles. This circuit can be added to any receiver so long as the tuned circuit is resonant at the i-f frequency and the lead from the tuned circuit to the i-f is not too long. Best results are obtained when the "Q" Multiplier is connected to the plate of the mixer feeding the lowest frequency i-f (in the case of dual conversion receivers).

The "Q" Multiplier is a simple device for obtaining additional receiver selectivity. The circuit for the Hammarlund HQ-160 "Q" Multiplier is illustrated in fig. 7.6-B.



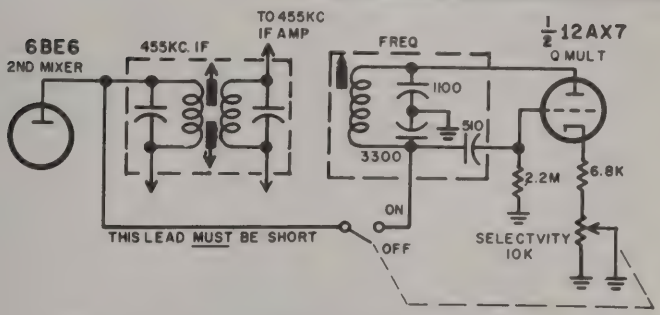


Fig. 7.6B—Schematic of the HQ-160 Q Multiplier. This device may be added to any communications receiver for increased selectivity.

The Drake 1-A triple conversion sideband receiver uses an interesting 50 kc filter at the last i-f frequency. Experimenters could construct such a filter, shown in fig. 7.6-C, by using high "Q" cup-core coils or small toroid forms. If the inductance was not adjustable, such as would be the case with toroids, it would be necessary to add or subtract turns. Each coil should be mounted in a compartment of its own, or mounted in such a manner so as to minimize the coupling.

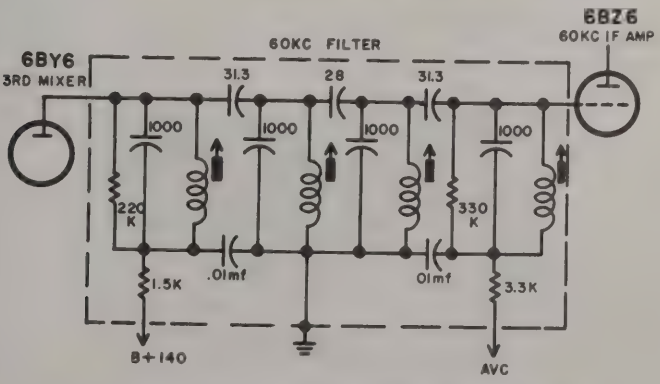
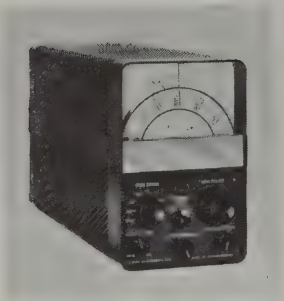


Fig. 7.6C—The Drake 1-A sideband receiver uses a 60 kc i-f filter. Experimenters could easily duplicate this filter.



The Drake 1-A sideband receiver uses a 60 kc last i-f. A simple 60 kc i-f filter is shown in fig. 7.6-C, and is used in this receiver for rejection of unwanted frequencies.

Interference Rejection

The so-called "notch" or "slot" filters are being used more-and-more in communications receivers for nulling out heterodynes. Again, these filters have been used in the better TV sets for elimination of adjacent channel and sound interference. Two popular forms are the bridged "T" trap, and the bifilar "T" trap.

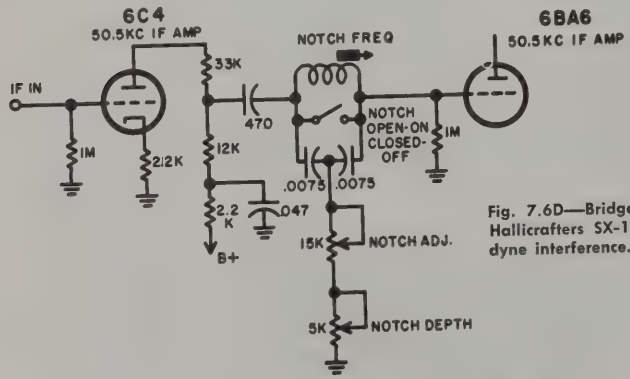


Fig. 7.6D—Bridge "T" trap used in the Hallicrafters SX-101 to eliminate heterodyne interference.

The bridged "T" is used in the 50 kc. i-f amplifier of the Hallicrafters SX-101, and is shown in fig. 7.6-D. The filter can be used at any i-f so long as the coil and series-connected capacitors resonate at that frequency. In fig. 7.6-D, the

The Hallicrafters SX-101 scuttles the heterodynes with an effective 50.5 kc "Bridged T" filter. The circuit diagram for this filter is shown in fig. 7.6-D.

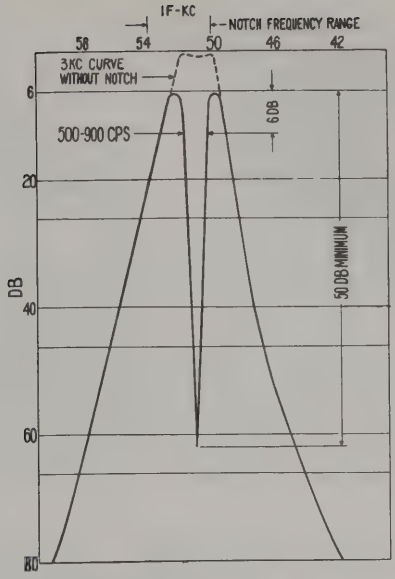


Fig. 7.6E—Heterodyne trap in the i-f amplifier places a sharp notch in the i-f response as shown. This curve is for the Hallicrafters SX-101.

inductance of the coil will determine the frequency of the "notch", and the 5K potentiometer will determine the depth. A minimum of 50 db is available in the SX-101, as shown in fig. 7.6-E.

The Hammarlund HQ-170 uses a bifilar "T" trap system. This slot filter will produce somewhat more rejection (about 60 db minimum) but is far more complicated and requires low input and output impedances. Fig. 7.6-F shows the bifilar

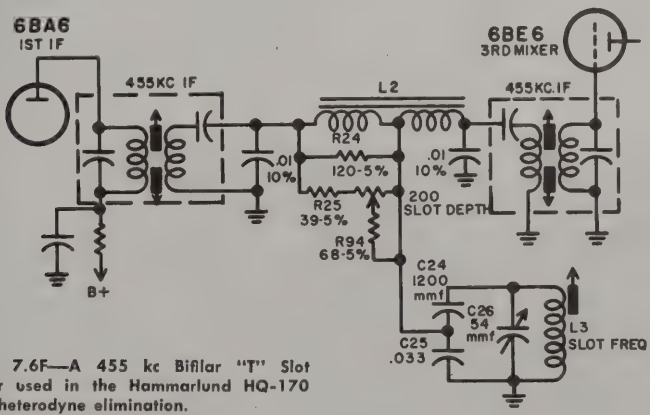


Fig. 7.6F—A 455 kc Bifilar "T" Slot filter used in the Hammarlund HQ-170 for heterodyne elimination.

Controls for the Bifilar "T" notch filter, discussed in the text, are located in the upper right hand corner of the Hammarlund HQ-170. They include slot depth and slot frequency.

"T" trap. Coil L2 is bifilar wound and broadly resonant at the i-f (455 kc). Resistors R24, R25, R26, and R94 determine the depth of the null, while C24, C25, C26, and L3 determine the position of the null with respect to the receiver i-f. Capacitors C24 and C25 place an "artificial tap" on coil L3, and their values are quite critical. Coil L3 should have as high a "Q" as possible. C26 and R26 are both adjustable from the front panel. The coupling i-f coils are standard units but note that the output and input winding (respectively) are series connected to obtain an impedance step down.

Product Detectors

Almost all SSB receivers use some form of product carrier detector to eliminate the evils that are present with the common diode detector. Two versions, either of which can be added to your communications receiver, are shown in fig. 7.6-G. Many hams think that the product detector is a panacea for SSB reception, but without stability and selectivity, the improvement may be masked by the other troubles. A product detector is just a fancy name for a converter or mixer. A similar device exists in your communications receiver in the 1st mixer stage. In the case of the product detector, the bfo is the local oscillator and the audio output is the i-f.

Fig. 7.6-G is a cathode follower type of product detector. Without the bfo signal it is nothing more than a cathode coupled class A amplifier, and therefore cannot detect. However, the level of the bfo signal is so large that it drives the sections from saturation to cut-off, thus carrying the tube into the non-linear (and mixing) region. If your bfo is not powerful enough to "cut the mustard" it may be necessary to add a buffer stage. Actually this is desirable since the bfo is more tightly coupled to the detector than in AM receivers.

The receiver i-f response can be shaped with ease by using any one of a variety of compact mechanical filters. This ingenious device uses mechanically resonant discs to filter unwanted frequencies.

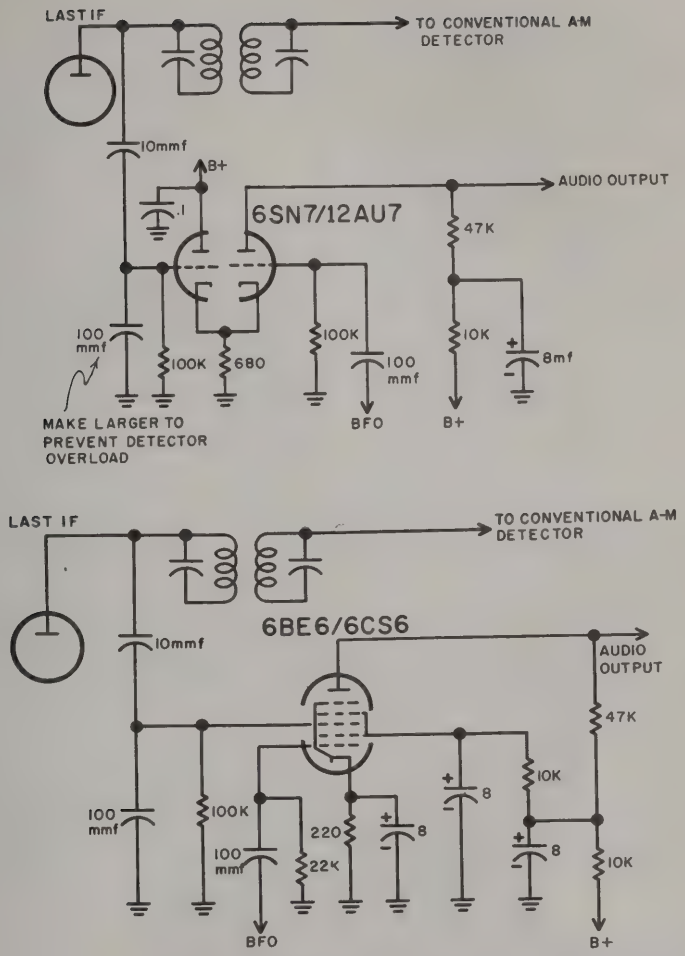


Fig. 7.6G—Two forms of "linear" detector. Either of these circuits is well suited for SSB demodulation.

There are two main advantages in using a product detector. The injection voltage is no longer critical, as it is with the diode detector and there will be less interference since output can only occur when a signal beats with the bfo. An excellent way to check the operation of the product detector is to disable the bfo. If the detector is not being overloaded, then the output from the product stage should cease. It is far more pleasant to copy AM stations with the product detector, since the bfo acts as a reinserted carrier, and the effects of phase distortion are minimized.

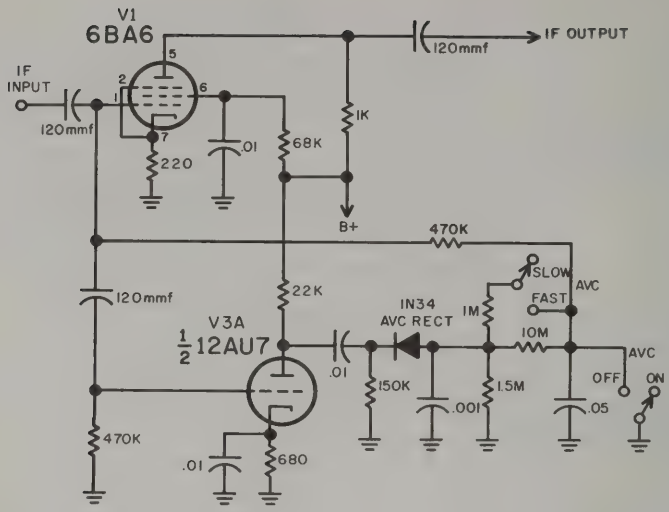
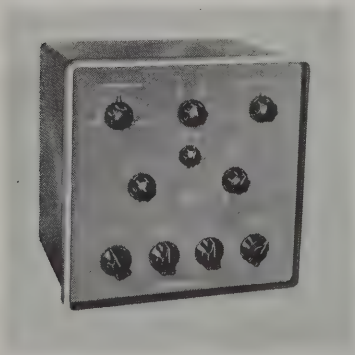


Fig. 7.6H—AVC section of the GSB-1. A 40 db change in input level will produce only 9 db output variation.

AVC

An extremely clever avc circuit that can be added to any adapter or i-f amplifier, is used in the Technical Materiel Corp. Model GSB-1 sideband adapter. Fig. 7.6-H is the diagram of the "bootstrap" avc system. R-F from the converter i-f transformer is applied to the grid of V1, the i-f amplifier and V3a, the avc amplifier. The 12AU7 amplifies the r-f and applies it to a 1N34 rectifier, connected to produce negative voltage. The detector output is tied to an audio filter, that has a selection of either long or short time constant components. This filtered negative voltage (that is in proportion to the signal level) is applied to the grid of V1 for gain control. The circuit is so effective that a 40 db variation of input signal will only produce a 9 db output change!



The Hammarlund HC-10, a post i-f adapter, contains more tubes and components than many communications receivers. Selectivity is obtained by using many high "Q" inductances in the i-f amplifier. A bifilar "T" notch trap in the input is employed to null out heterodynes (see section 7.6).

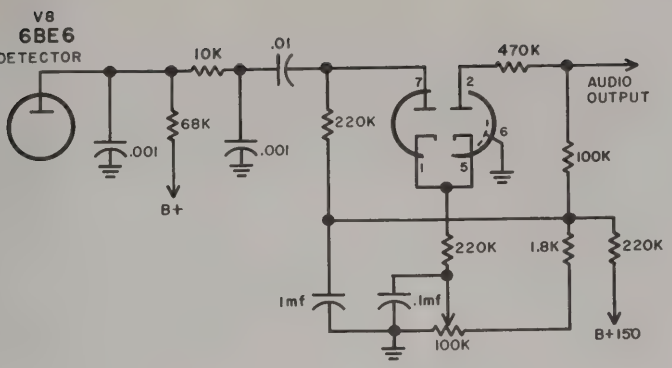
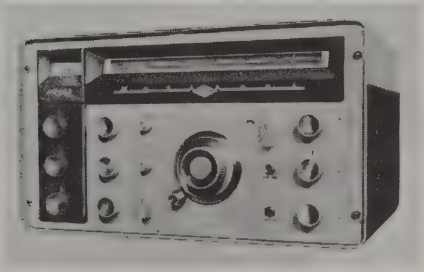


Fig. 7.6I—NC-303 CW-SSB noise limiter circuit.

SSB Noise Limiter

Noise problems have always been a thorn in the side of SSB operators. Methods of minimizing such noise have been very ineffective, until the introduction of the NC-303. This receiver uses a noise limiter, illustrated in fig. 7.6-I, that is said to be very effective for SSB or CW reception. The output of the product detector is applied to a 6AL5 dual-diode, through a filter network. The clipping level may be varied by changing the cathode voltage with the 100K potentiometer. Audio output is taken from the pin 2 section of the 6AL5. The clipper is left in the circuit at all times. Moving the arm of the potentiometer to the ground end eliminates the clipping.



The National NC-303 contains a SSB/CW noise limiter which is described in section 7.6.

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8

Accessories

You can use your Collins receiver as a VFO with the P&H "VFO-Matic". This clever device modifies the receiver local oscillator signal and converts it to the proper injection frequency for such exciters as the Multiphase, Lakeshore Phasemaster, KWS-1, or HT-32, making the transmitter "track" right along with the receiver.

I HAVE no doubt that an entire book could be written about the various accessories that are available for the sideband station. By the same token, no book on sideband techniques could be considered complete without mention of some of the more popular items, such as variable frequency oscillators, voice operated relays, phone patches, and electronic antenna relays.

The Lakeshore "Time-Master" is a handy accessory for any station. It will remind you when those 10 minute station identifications are required.

8.1 VARIABLE FREQUENCY OSCILLATORS

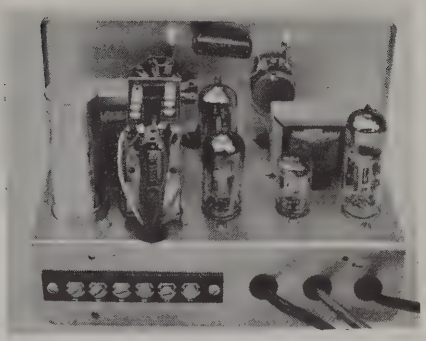
Many SSB transmitters and exciters contain a variable frequency oscillator (vfo) for shifting the output frequency, while others such as the Central Electronics 10A and 10B require an accessory vfo. A BC-458 Command Set transmitter (5.3- 7.0 mc) makes an ideal unit for this application. When properly converted this vfo will have stability in the order of 100 cycles or less. The usual conversion consists of removing the rolling coil and one 1625. A B plus voltage of 75 volts, regulated, is supplied to the oscillator and the plate and screen of the 1625. Several companies recommend using dc for the filaments to eliminate hum modulation, but the author has never found this necessary. It is a good idea to provide for 24 hour operation of the filaments to keep the BC-458 near operating temperature.

A few thousand well chosen words on the circuits and peculiarities of vfo's would be in order, for this is a critical part of the SSB station.

Stability

Frequency stability of oscillators used for carrier generation and re-insertion is a prime requirement. The frequency stability required for an AM transmitter is not as stringent, and often two or three kc of drift is tolerated. The unhappy person who attempts to convert such a transmitter to DSB or SSB soon finds the situation hopeless. It can be an extremely difficult task to rework a vfo to produce only a few hundred cycles of drift. If you have ever dug around in the bowels of a vfo, you may have asked yourself "Why does it drift?" Well, why does it drift? To understand the reason for this annoying defect of many transmitters, let's explore the inner secrets of the variable frequency oscillator.

Drift can be separated into two distinct categories, i.e. mechanical and electrical. Mechanical drift is the result of poor design and it is usually caused by insufficient rigidity of the chassis or oscillator components, or both. It is truly amazing to note how much havoc the slightest movement of a wire, in the oscillator circuit, can cause. If the chassis is tensing the wire, the frequency can shift wildly simply by moving the cabinet or chassis. This form of instability can be minimized by lashing all loose wire and components securely, and by "beefing up" the chassis in critical spots. It goes without saying that any inexpensive components should be replaced with high quality units. As an example, compression padders should be replaced with rotary disc or air padder capacitors. Slug tuned coils are also a source of mechanical instability. The coil form tolerances may allow the slug to flop around, thereby causing inductance variations. The cores should be secured with wax whenever an oscillator is aligned or when looking for mechanical instability. Never cement the slugs, for when it dries, the cement will "pull up" and change the position of the slug.



The VX-1 has a self contained power supply. The relay uses extra heavy contacts for switching inductive loads.

Electrical drift is somewhat more elusive. This form of instability is the result of temperature variations which cause the components to change value. Coils, capacitors, vacuum tubes, in fact almost any component will change value with a change in temperature, some more than others. If the component is in a frequency determining part of the oscillator circuit, a change in operating frequency will occur. Should the oscillator frequency drift for a few minutes and then "settle down", it is said to exhibit *short term drift*. A critical frequency determining component mounted near a vacuum tube might cause short term drift. *A long term drift* might be characterized by an oscillator frequency that continues to drift for hours. This could be caused by mounting a frequency determining component on the chassis. Whereas the vacuum tube comes up to operating temperature quite rapidly, it may take many hours for the chassis to "warm up".

Temperature drift is inherent in any oscillator circuit, but the effects may be subdued in several ways. The first consideration in a stable oscillator is to use the finest components available. The small temperature drift of these components may then be corrected, or compensated, by incorporating temperature compensating capacitors. In effect these capacitors introduce additional drift that is "180° out-of-phase". The capacitors have a controlled temperature versus capacity range and are produced by such companies as Centralab, Erie, and many others. By combining, in proper amounts, zero and negative temperature coefficient capacitors, it is possible to counteract component drift almost 100%.

To determine if you are "making any headway" it is necessary to measure the amount of drift in a particular period of time. A set-up for making such measurements is shown in fig. 8.1-A. The variable frequency oscillator is adjusted for zero beat with a stabilized crystal oscillator and monitored on a communications receiver.

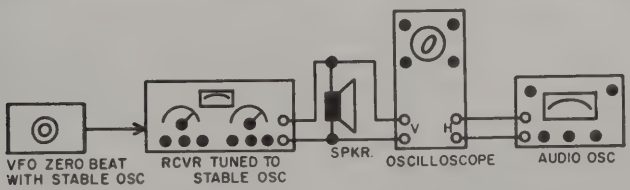


Fig. 8-1A—Equipment set-up to accurately measure frequency drift.

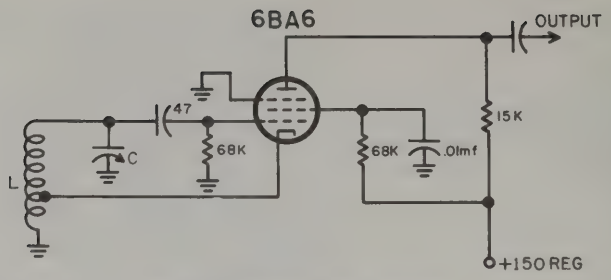


Fig. 8.1B—Electron coupled oscillator.

The beat difference (as heard in the speaker) is fed to the vertical input of an oscilloscope. A signal from an accurate and stable audio oscillator is then applied to the horizontal input. A circular Lissajou pattern will be seen when the audio oscillator is the same frequency as the beat note. By reading the frequency every few seconds, it is possible to draw a very accurate graph of frequency versus time.

Temperature drift problems may be minimized by the proper selection of oscillator circuits also. The two most popular oscillator circuits, presently being used, are shown in fig. 8.1-B and 8.1-C, and are the electron coupled oscillator (eco) and the Clapp oscillator. It should be pointed out that the degree of

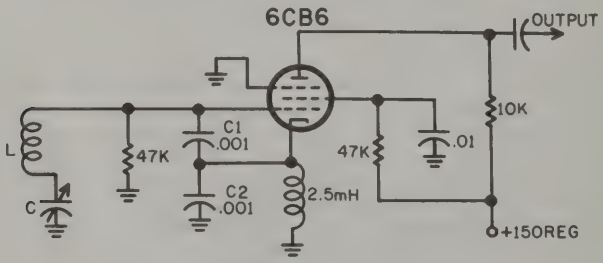


Fig. 8.1C—Clapp oscillator.

coupling between the tank circuit and the associated vacuum tube will be a large determining factor in the stability of the oscillator circuit. If the oscillator circuit consisted of nothing more than a coil and a variable capacitor, the only source of drift would be due to temperature, and this could be easily corrected with temperature compensating capacitors. Unfortunately, in order to sustain oscillation, the tank circuit must be connected to a vacuum tube circuit capable of amplification. This is where the trouble starts. Not only is the vacuum tube subject to temperature and capacity variations, but the impedance presented to the tank coil can vary widely.

The electron coupled oscillator is used by Collins Radio Company in their 70E and H series of permeability tuned oscillators. A typical circuit of such an oscillator is shown in fig. 8.1-E. Although the tank coil is tightly coupled to the oscillator through a 47 mmfd. capacitor (and the cathode tap on the coil), the unit is so well compensated that the electrical drift is usually less than 400 cycles. The mechanical drift is non-existent because of its rugged construction.

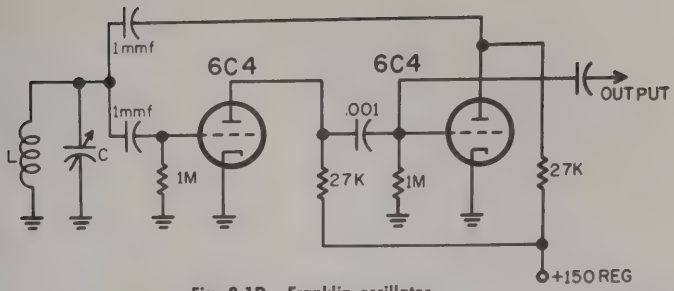


Fig. 8.1D—Franklin oscillator.

In the electron coupled oscillator, the feedback voltage appears across the tapped portion of the coil. It is inductively coupled to the portion of the coil above the tap and applied to the control grid. The amplified signal again appears across the tapped portion of the coil, and a feedback path is created. Thus, the stage continues to oscillate. Output from this circuit is obtained from the plate circuit and no connection, other than the electron stream, exists between the output and the tank circuit. Variations of the output load will have a minor effect on the oscillator frequency. By proper selection of the plate and screen voltage dropping resistors it is possible to make this circuit virtually independent of supply voltage.

The Clapp oscillator (shown in fig. 8.1-C) is considered to be inherently more stable than the electron coupled oscillator. This circuit is easily recognized by the fact that the tank coil is series tuned. The tank circuit consists of the coil "L", tuning capacitor "C", and the divider capacitors C1 and C2. These capacitors are made as large as possible while maintaining stable oscillation. The reactance of C1 and C2 is usually about 100 ohms at the operating frequency and therefore, the oscillator tube is said to be coupled to a low impedance portion of the tank circuit. The net result is much the same as connecting the tube near the cold end of the coil. Variations in the characteristics of the oscillator tube will have a minimum of effect on the tank circuit. At resonance, high current flows in the capacitors and the voltage appearing across C1 is fed to the input of the oscillator. The amplified output appears across C2 as feedback voltage and the stage continues to oscillate. The output circuit is the same as the eco and operated in a similar manner. A tuned circuit or an r-f choke can be substituted for the plate load resistor for more

The Collins 70E permeability tuned oscillator uses the electron coupled circuit plus a buffer stage and exhibits a high degree of stability. The unit is hermetically sealed to prevent component deterioration.

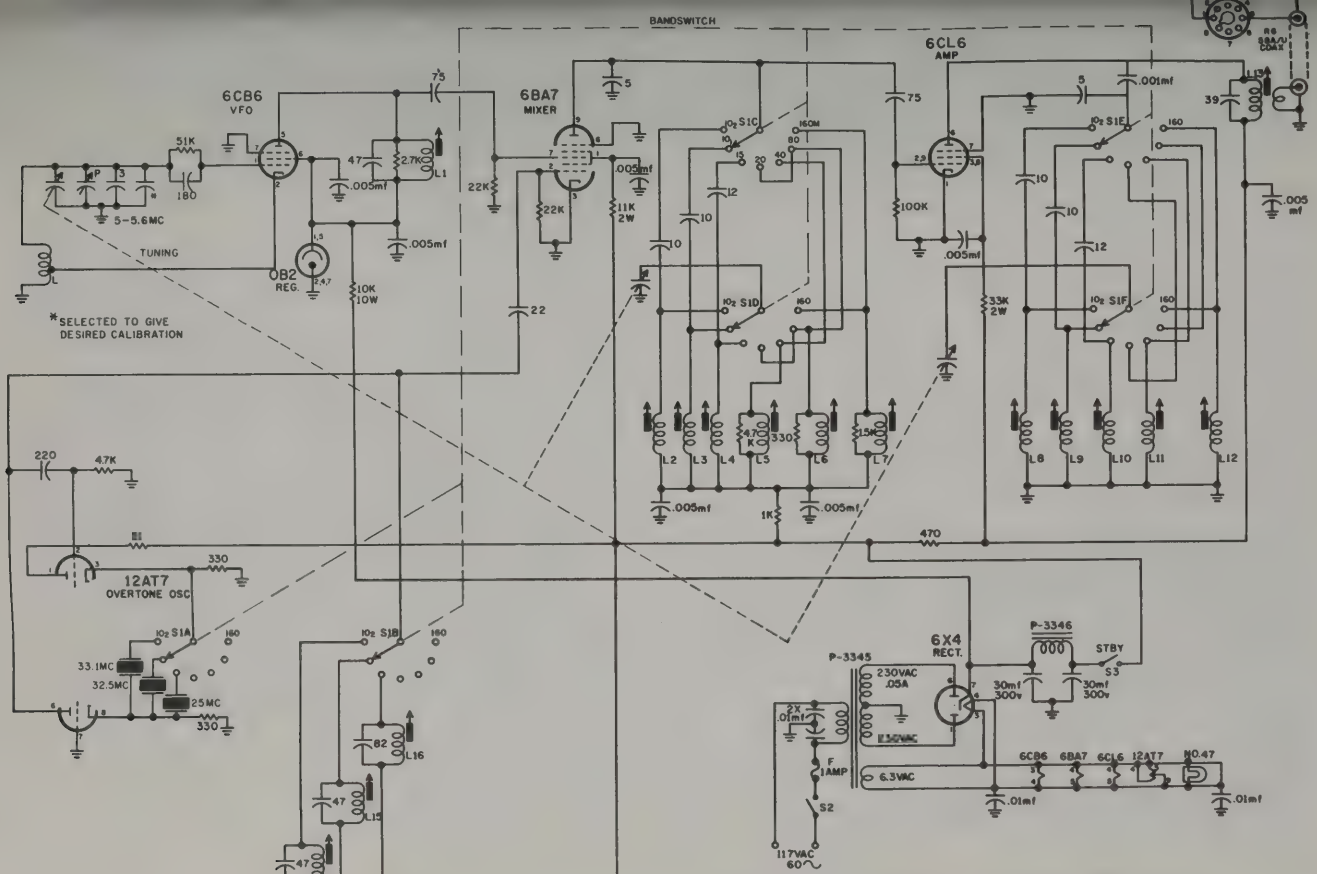


Fig. 8.1G—Schematic diagram for the Lakeshore "Bandhopper" vfo. This unit provides the proper injection frequency for all band operation of the popular 9 mc phasing exciters.

The 70K-1 is the master oscillator for the Collins KWM-1. It uses a single 5749 (rugged 6BA6) in a Clapp oscillator circuit.

output, however, the oscillator stability will be reduced. Although the name Clapp (the inventer) has become popular for this circuit, it is truely a series tuned electron coupled oscillator.

A practical application for the Clapp oscillator is shown in fig. 8.1-F. This is the circuit of the Model 70K-1 permeability tuned vfo used in the Collins KWM-1. Coil L24 is ganged to the knob on the front panel and L23 is used to limit the range of inductance. Capacitor C147 is selected by the manufacturer for the proper degree of temperature compensation. The calibration may be corrected by the adjustment of C148. The dividing network consists of C144 and C145. The output signal appears across R129 and is coupled to the associated circuits via C141. This oscillator is designed for use in mobile equipment and exhibits less than 50 cycles of mechanical drift when subjected to a 5G shock. A typical warm-up drift is less than 200 cycles.

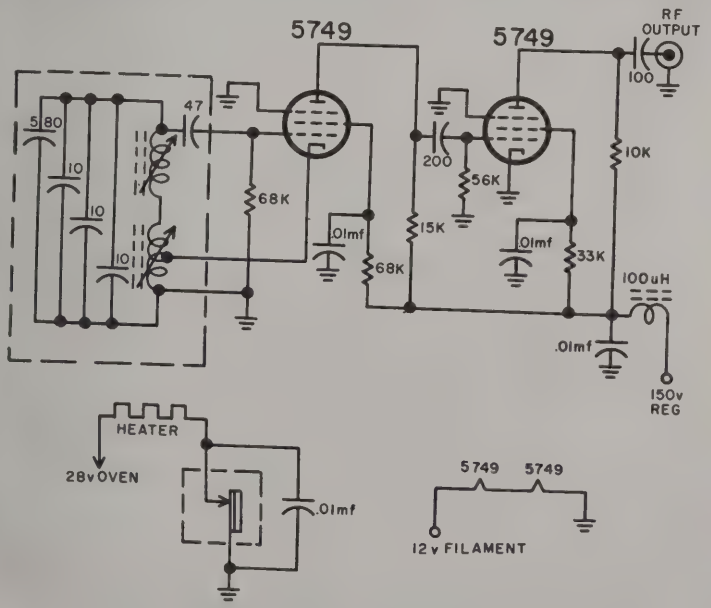
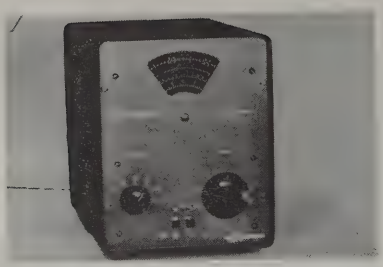


Fig. 8.1E—Circuit diagram of a typical Collins oscillator, similar to the type used in the 75A receiver and 32V transmitter series.



The Lakeshore "Bandhopper" VFO uses many of the high quality Command Set transmitter components in an oscillator circuit operating between 5.0 and 5.6 mc. The electron coupled oscillator output either operates straight through, is multiplied, or is mixed with the heterodyne oscillator.

The author's favorite oscillator is the "old time" Franklin, and it is believed to be the most stable of them all! This rock solid device can put a quartz crystal to shame! Because it represents the ultimate in stability, it makes an ideal vfo for sideband applications. The basic Franklin oscillator is shown in fig. 8.1-D, and a practical adaption is described in section 8.2. This oscillator differs from the others in that the tank circuit is in *shunt* with the feedback path, while the tank in the eco and Clapp is in *series* with the feedback. The tuned circuit is very lightly coupled to the oscillator (typical value 1 mmfd.). Two cascaded tubes (usually a dual triode) provide the necessary high transconductance necessary to maintain oscillation. If you mentally remove the tank coil and capacitor from the circuit, it has all the appearances of a multivibrator and, in-fact, that is exactly what it is. However, with the tuned circuit shunting the feedback path, the only oscillation that can occur is at the natural resonant frequency of the coil and capacitor. The output from this oscillator is very low. It is necessary to incorporate a tuned buffer circuit even for the low drive requirements of a local oscillator. To obtain the customary 5- 10 volts necessary for transmitter drive, the buffer stage should be untuned, and be followed by a tuned amplifier.

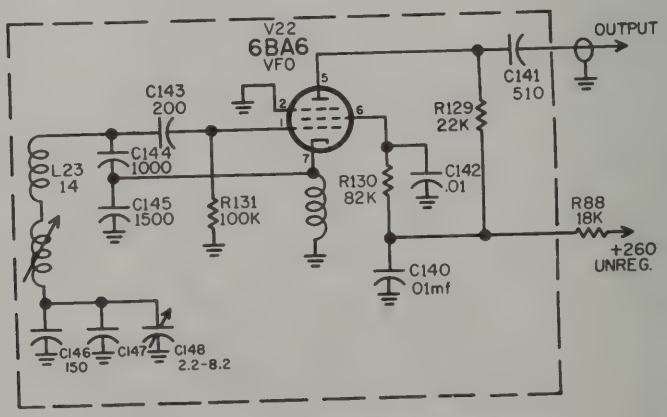


Fig. 8.1F—Typical application of the Clapp oscillator circuit in the portion of the KWM-1 transceiver.

8.2 A DRIFTLESS VFO

The circuit shown in fig. 8.2-A illustrates a practical application of the basic Franklin circuit. It would be difficult for the average ham to duplicate a Collins v.f.o. even if an elaborate machine shop was available. It is safe to say that the "Driftless VFO" is the most stable vfo that a ham with average facilities can construct. The mechanical stability will be proportional to the workmanship of the constructor. The electrical stability of the "Driftless VFO" will be *less than 200 cycles* and occurs in the *first 30 seconds of operation!* This should be repeatable if the constructor uses the same components and the same chassis layout. It should also be noted that zero temperature coefficient capacitors are used in the oscillator section, and no form of temperature compensation is employed. Truely, the stability is only limited by the care taken in the mechanical construction of the vfo.

The "heart" of the vfo is the tuning capacitor. It was removed from an old derelict Command Set transmitter that had seen better days. Originally, it was used to resonate the final amplifier and contains the gear mechanism for the dial. Did you ever take a close look at this capacitor? If not, do so. It is one of the most beautifully constructed variable capacitors that can be found in war surplus equipment. The gears incorporate a *working* anti-backlash system. The rotor plates "float" in ball bearing end caps, complete with tension adjustments. The stator plates are held firmly in place with glass ball insulators, also complete with tension adjustments! Needless to say, it makes an excellent capacitor for the high stability vfo. If you purchase a Command Set transmitter to obtain this capacitor, examine it carefully for damage. Make sure that the insulators or bearings have not fallen out.

The tank inductance consists of a slug tuned coil manufactured by National (XR-50). The slug is held firmly in place with a spring loaded mounting. The slug does not appear to move, and no mechanical instability can be blamed on this coil. The final component in the tank circuit "team" is a variable air padder. This capacitor is used to track the high frequency end of the oscillator range, while the core of the XR-50 trims the low frequency end. The tank circuit is coupled to the oscillator with 2.2 mmf tubular NPO capacitors. The oscillator tube V1 is a type 12AV7. All of the common dual triodes (12AT7, 12AU7, 12AY7, etc.) were tried in this circuit and the type selected was found to be superior. The oscillator is directly coupled to the broad-band buffer stage, and this in turn drives the output 6CB6. In addition to stepping the r-f up to a usable level, the high degree of isolation precludes the possibility of frequency shift due to load variations. The power supply is a half wave circuit incorporating a silicon rectifier and plenty of filtering. A two section wafer switch is used to energize the filaments during standby periods. The other section energizes the B plus buss.

Construction

The vfo is completely contained in a cabinet measuring 4″ x 7″ x 9″. These cabinets have been designed expressly for Amateur equipment, to provide a commercial appearance. The body of the cabinet is perforated steel with steel end caps. The aluminum chassis secures the end caps with weld nuts. This results in an extremely rugged assembly. These grey hammertone cabinets are available from California Chassis Company, 5455 Century Blvd., Lynwood, California. Their Model LTC-463 was used in the construction of this vfo.

The Command Set transmitter tuning capacitor (C1) was mounted in a rather unusual manner. A 3″ x 4″ cut-out was made at the front of the chassis to recess the variable capacitor. Unless this is done, the tuning knob will protrude very close to the top of the cabinet. By sub-mounting one end of the capacitor, it allows the knob to be positioned at the intersection of the dial and the pilot lamp holder. It does create an extra bit of work, though, and it is not really necessary for correct

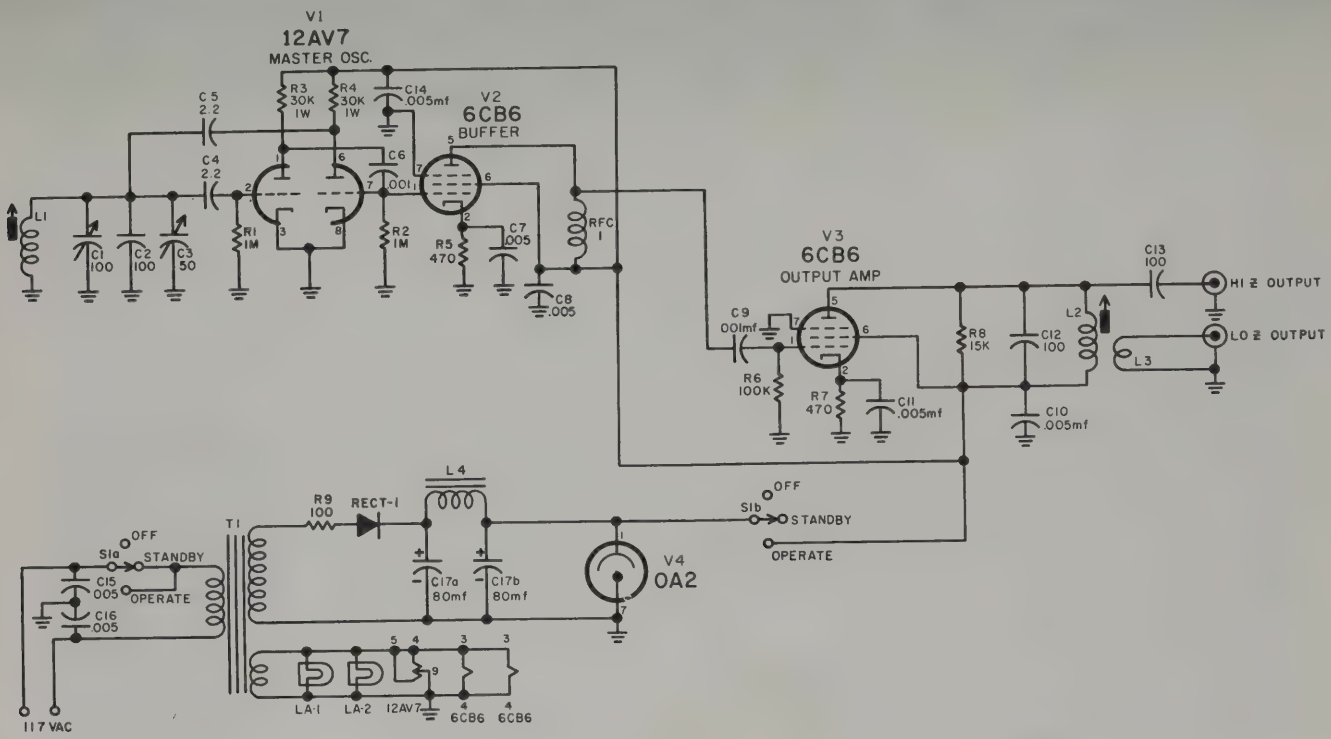
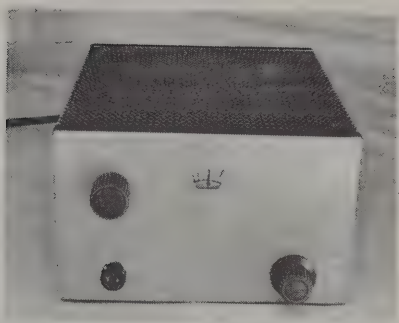


Fig. 8.2A—A driftless vfo.



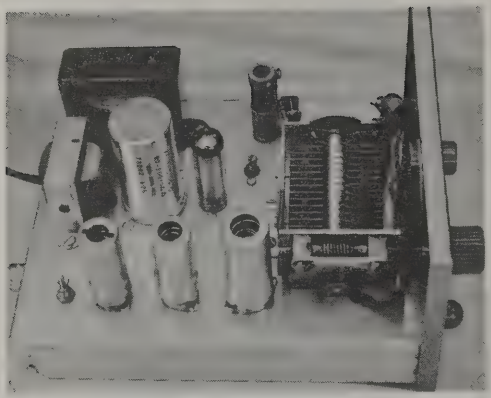
Presenting the Driftless VFO. This device has less than 50 cycles drift after a 1 minute warm-up.

operation of the vfo. The coil, L1, was mounted at the side of the tuning capacitor, away from the heat generating components. Trimmer capacitor, C3, was mounted directly behind C1. For mechanical rigidity, these three components were wired with #14 tinned bus-bar wire. An insulated stand-off, mounted near V1, is used to secure the oscillator coupling capacitors C4 and C5. The components associated with V1 should be grouped tightly around the tube socket, but not too near the chassis. A short straight wire between V1 and the control grid pin 1 of the buffer is essential. The output stage should be kept as far away from the oscillator as possible to prevent unwanted feedback. The layout of the rear apron is left up to the constructor, and will probably be determined by the type of connectors that are used. The pilot lamp holder is mounted under the chassis and on the side opposite the coil, L1. The leads from S1 are cabled together and routed down the side of the chassis away from the r-f circuitry.

The vfo has been used on several frequencies up to 7 mc. If it is desired to generate a stable signal higher than this frequency, the oscillator should be designed for half frequency and the output stage tuned as a doubler. The parts list includes coil and capacity data for both 3.5- 4.0 mc for use with transmitters employing fundamental frequency sideband adapters, or 5.0- 6.0 mc for exciters such as the Central Electronics and Lakeshore type. The output inductance (L2) is designed for either high or low impedance output. Because of the wide variety of capacities the vfo will be connected to, it may be necessary to juggle the size of C12, L2, or both. If maximum output occurs with the slug full inserted into L2, increase the size of C12. If the coil peaks with the slug all the way out, reduce the size of C12. Resistor R8 controls the output level of the vfo and the bandwidth of the output coil. With the values shown, the output voltage will be approximately 10 volts across the link and the bandwidth will be roughly 300 kc at 6 db down. If more output is required, resistor R8 can be eliminated and a tuning capacitor substituted for C12.

Parts List

C1—Main tuning — 100 mmfd. final amplifier capacitor from Command Set Transmitter
C2—100 mmfd. tubular n.p.o. (Centralab TCZ-100)
C3—50 mmfd. variable air padder (E. F. Johnson 50K10)
C4, C5—2.2 mmfd. tubular n.p.o. (Centralab TCZ-2R2)
C6, C9—.001 mfd., silver mica
C7, C8, C10, C11, C14, C15, C16—.005 mfd., 600 volt disc ceramic
C17—2 x 80 mfd., 300 WVDC (Sprague 1289)
L1—30 turns #26 enameled wire closewound on National XR-50 coil form (3.5- 4.0 mc.) 22 turns #24 enameled wire closewound on same type form (5.0- 6.0 mc.)
L2—14- 31 uh. scramble wound coil such as Miller 4407 (3.5- 4.0 mc.) Use same coil, but delete C12 for 5.0- 6.0 mc.
L3—10 turns #22 wound over the top of L2
L4—7 henry, 50 ma. filter choke (Stancor 1277 or Triad C-3X)
LA1, LA2—two #47 pilot lamps
R1, R2—1 meg., ½ watt
R3, R4—30K, 1 watt
R5, R7—470 ohm, ½ watt
R6—100K, ½ watt
R8—15K, ½ watt
R9—100 ohm, 2 watt
RFC—2.5 mh., powdered iron core r.f. choke (J. W. Miller #6302)
Rect. 1—Silicon rectifier, Audio Devices A-750 or Sarkes M-500
S1—two pole, three position wafer switch (Centralab 1472)
T1—125 volt, 50 ma. power transformer (Stancor PA-8421 or Triad R-30X)
Vacuum tubes—12AV7, (2) 6CB6, OA2

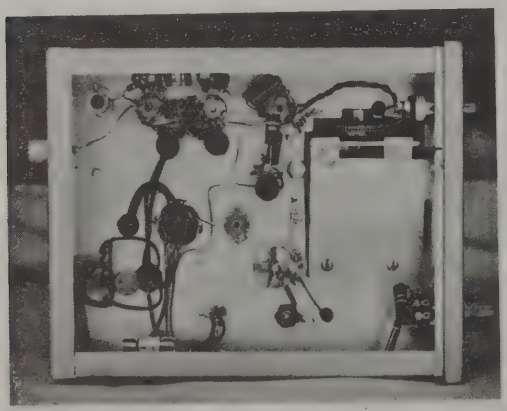


Inside view of the Driftless VFO. The coil is located on the chassis top, opposite the vacuum tubes. The variable capacitor is sub-mounted to lower the tuning knob. Note the number of plates in the variable capacitor, when used to tune the receiver in the KWM-1 (see section 8.3).

Alignment

The values for C4 and C5 are about optimum, although either one of these capacitors is often made an adjustable element. If the capacity is insufficient, the oscillator will not take off. If the capacity is excessive, the oscillator will start a blocking action akin to superregeneration. When checking out the vfo, check for a pure note, and tune on each side of the carrier to check for "birdies".

The high and low frequency end of the oscillator range is trimmed by adjusting capacitor C3 and the slug of L1 respectively. If only a limited tuning range is desired (as for the KWM-1 DX Adapter to be described) it will be necessary to remove some of the plates from C1. The desired frequency range should be spread over as much of the dial as possible to slow down the tuning rate. With the gearing of the capacitor, tuning is not critical even with a half megacycle range. Once the desired tuning limits have been set, coil L2 should be tuned to the center of the range with the vfo connected to the equipment that it is to drive. Since the input and cable capacity will affect the coil, the cable should be as short as possible if the high impedance output is used. A maximum length of three feet is recommended on high impedance, but any convenient length will be satisfactory, when using low impedance output.



The oscillator components are wired with # 14 tinned bus-bar wire. The modifications described in section 8.3 were made after this photograph was taken.

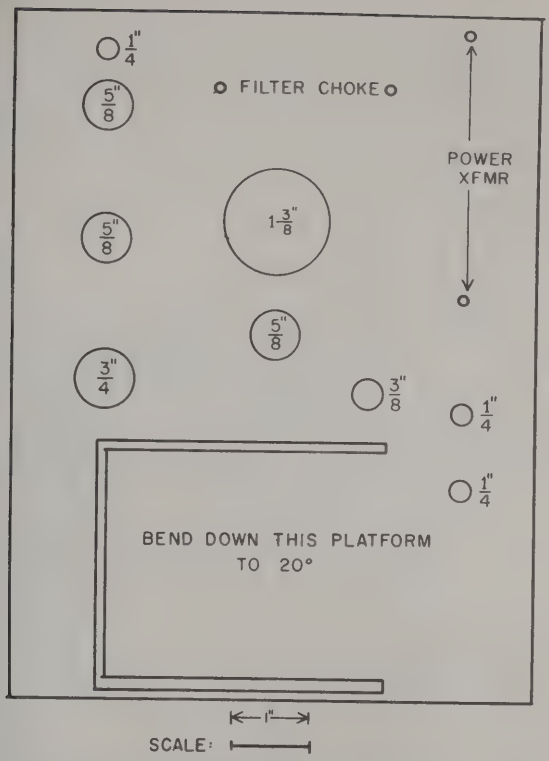


Fig. 8.28—Chassis layout for the driftless VFO.

8.3 A DX ADAPTER FOR THE KWM-1 TRANSCEIVER

The receiver portion of the KWM-1 is essentially a double conversion superhet. The first conversion oscillator is crystal controlled and is four megacycles above the incoming signal frequency. This signal is heterodyned into a variable i-f stage which tunes 3.9- 4.0 mc. The vfo for this stage is also used when transmitting and therefore the device sends and receives on the same frequency. This is a wonderful feature most of the time for it eliminates several fussy tuning adjustments. But when the DX starts rolling in (usually outside the American phone band) it can be a nuisance. Although the Collins Radio Company has available a DX adapter, it places the operator on a crystal frequency inside the band. For maximum flexibility, another 70K-1 vfo could be switched into the circuit when transmitting.

The terrific stability and slow tuning rate of the vfo just described makes it a "natural" for tuning the KWM-1 receiver. With this thought in mind, a conversion for the KWM-1, to separate the receiver and transmitter tuning, was worked out. For those of you who shudder at the thought of modifying a piece of commercial gear — relax! No holes are necessary, operation returns to normal simply by unplugging the vfo, and when you sell it, the wiring can be removed in 15 minutes and no one is the wiser.

The conversion uses an ingenious switching circuit to interchange vfo's. When power is applied to the external vfo (by switching to operate) the added relay will actuate and connect the external vfo signal to variable i-f mixer tube. However, when you talk and trip the vox circuit, the relay automatically drops out, thereby reconnecting the internal vfo. Clever, what? The modifications to the KWM-1 are shown in fig. 8.3-A.

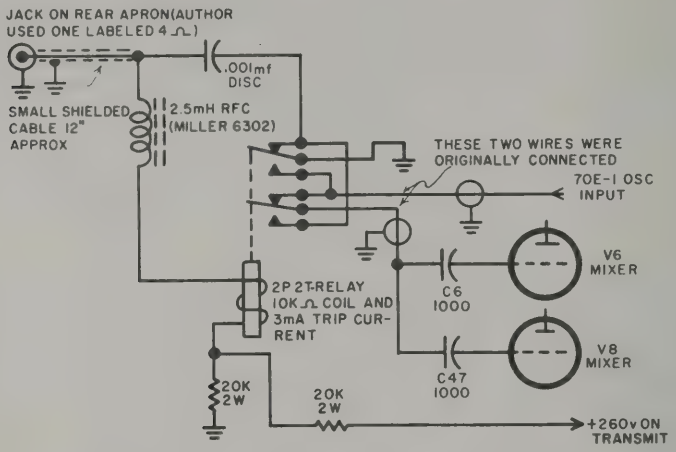


Fig. 8.3A—These simple changes will modify the KWM-1 for completely independent receiver and transmitter tuning.

The KWM-1 vfo operates 455 kc below the variable i-f and tunes 3.445 mc to 3.545 mc. To receive, say 14 mc, an 18 mc crystal is switched into the circuit, and a beat difference of 4 mc occurs. This 4 mc signal is again heterodyned with a 3.545 mc signal from the vfo and the difference frequency of 455 kc is amplified in the i-f. Signals on 14.1 mc may also be received in this crystal position. The 14.1 mc energy beats with the 18 mc crystal oscillator and produces a difference of 3.9 mc. Moving the vfo frequency to 3.445 mc allows this signal to be heterodyned into the i-f channel. Therefore, a vfo frequency of 3.545 mc corresponds to the crystal frequency plus 100 kilocycles.

The remote VFO relay is located in the space between the coil shields. These modifications, described in the text, can be made in the KWM-1 without drilling any holes in the equipment.

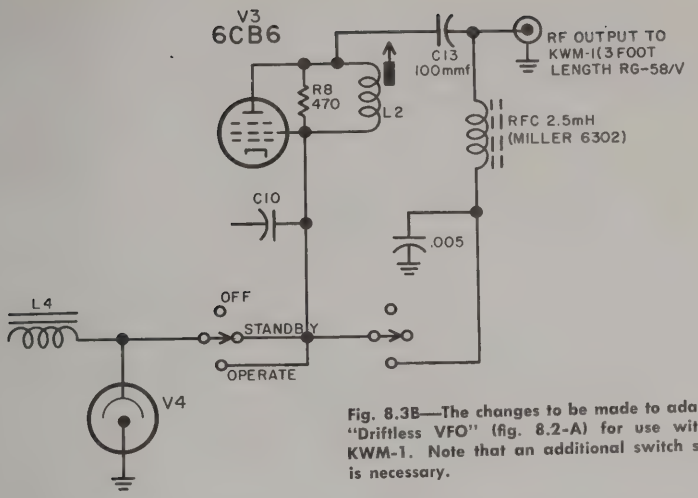


Fig. 8.3B—The changes to be made to adapt the "Driftless VFO" (fig. 8.2-A) for use with the KWM-1. Note that an additional switch section is necessary.

It is necessary to change the vfo circuit slightly, as shown in fig. 8.2-A. The vfo is designed to tune plus or minus 200 kc from the crystal frequency. From the preceding explanation, it can be seen that the vfo tank circuit must be padded to cover the range of 3.345 to 3.745 mc. In addition dc voltage must be applied to the coax to trip the remote vfo relay. This is accomplished by letting the vfo run continuously and applying dc to the cable (through an r-f choke) when remote tuning is desired. Complete the modifications to the "Driftless VFO" by removing C12 and changing R8 to a 470 ohm, ½ watt. This will lower the drive to the KWM-1 mixer. Coil L2 will have little (if any) effect on the output. These modifications are shown in fig. 8.3-B.

The KWM-1 circuit modifications are shown in fig. 8.3-A while the location of the components can be seen in the accompanying photographs. To illustrate how this novel switching circuit works, assume that nothing is connected to the input jack. Signals from the 70K-1 vto in the KWM-1 pass through the upper relay contacts and on to the two mixer tubes. When the transmitter is energized approximately 150 volts is applied to the bottom end of the relay but it fails to trip since the top end of the relay circuit is incomplete. In other words, with no connection to the vfo input, the KWM-1 operates normally (that is, transmits and receives on the same frequency). Now, let's chase DX by plugging in the remote vfo. Remember it was modified earlier to apply 150 volts to the cable whenever the vfo was energized. This 150 volts flows through the r-f choke (which opposes the r-f from the vfo), through the relay and 20K resistor back to ground, thereby tripping the relay. This changes the contact so that the remote vfo signal passes through the .001 mf capacitor and on to the mixer tubes (V6 and V8). On transmit, the voltage applied to the bottom end of the relay (from the KWM-1 B plus circuit) equalizes the voltage across the relay, and it "drops out". This connects the mixer tubes to the internal vfo. Thus, whenever the remote vfo is plugged in and energized, it will tune the receiver plus or minus 200 kc each side of the crystal frequency. When transmitting, the relay is de-energized and the internal vfo supplies the signal. An additional set of contacts on the relay is used to short out the unused signal source. If this is not done, you will find yourself transmitting two carriers! The unused vfo will only be 20- 30 db weaker than the main signal. The FCC frowns on this!

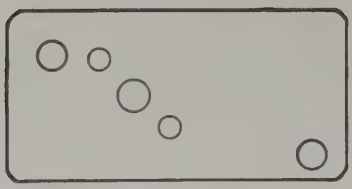


Fig. 8.3C—Drilling template for mounting VFO relay. The center holes should be drilled to suit relay.

Modifications

The relay for switching vfo's was installed in an empty space between the coil shields. A hole in the KWM-1 chassis was covered up by a dummy plate covered up by two screws. This plate can be drilled to secure the relay or another plate can be made and set above the original by ¼ inch spacers. A template for such a plate is shown in fig. 8.3-C. You will note a very tiny white shielded wire passing through this space between the coil shields. Trace the wire back to its source. It should come from the 70K-1 vfo unit and the other end should be tied to a terminal strip near the mixer tubes. Clip this wire at the relay, prepare the two ends and connect to the relay as shown in fig. 8.3-A. Wire the rest of the circuit making sure that none of the components will short to the bottom when it is replaced. The source of plus 260 on transmit is located on a terminal lug on the relay adjacent the newly added vfo relay. A red- white- blue is connected to the 260 volt terminal. It is recommended that you check the voltage at this point, just in case Collins might have changed the color code. Complete the modifications by preparing a 12 inch length of *small* coax or shielded cable. Select an unused jack on the rear apron of the KWM-1 and remove and tape the wire. This should be relabeled VFO INPUT. Connect one end of the cable to this jack and the relay end to the junction of the choke and capacitor. A dial can be made for the remote vfo which reads plus 200- plus 100- 0- minus 100- minus 200, with calibration marks at the appropriate points. Happy DX chasing with the KWM-1 DX adapter.

8.4 VOICE OPERATED RELAYS

One of the many features that attract hams to the SSB mode of transmission, is voice actuated transmissions. This system never became very popular with AM transmitters, but is considered the rule rather than the exception on sideband. The idea of VOX operation is simplicity itself. The microphone voltage (after suitable amplification) is applied to a diode rectifier. The output is positive in polarity and this dc voltage is applied to the grid of a relay control tube. The "voice voltage" overcomes the bias on the control tube and trips a plate current relay. The contacts on the relay are used to switch over the various functions in the station equipment.

One might think that sounds emitted from the receiver speaker would trip the VOX. A clever circuit is included in the VOX system to eliminate this possibility. Audio from the speaker output is applied to a one stage amplifier and then rectified by another diode section. The output polarity of this section, however, is negative, and is also applied to the relay control tube. Thus, any sound picked up by the microphone from the speaker produces both a positive and a negative voltage at the grid of the control tube and therefore cancels out.

An interesting voice operated relay is the Heath VX-1 and is shown in fig. 8.4-A. Audio from the microphone is applied to a speech amplifier stage using a

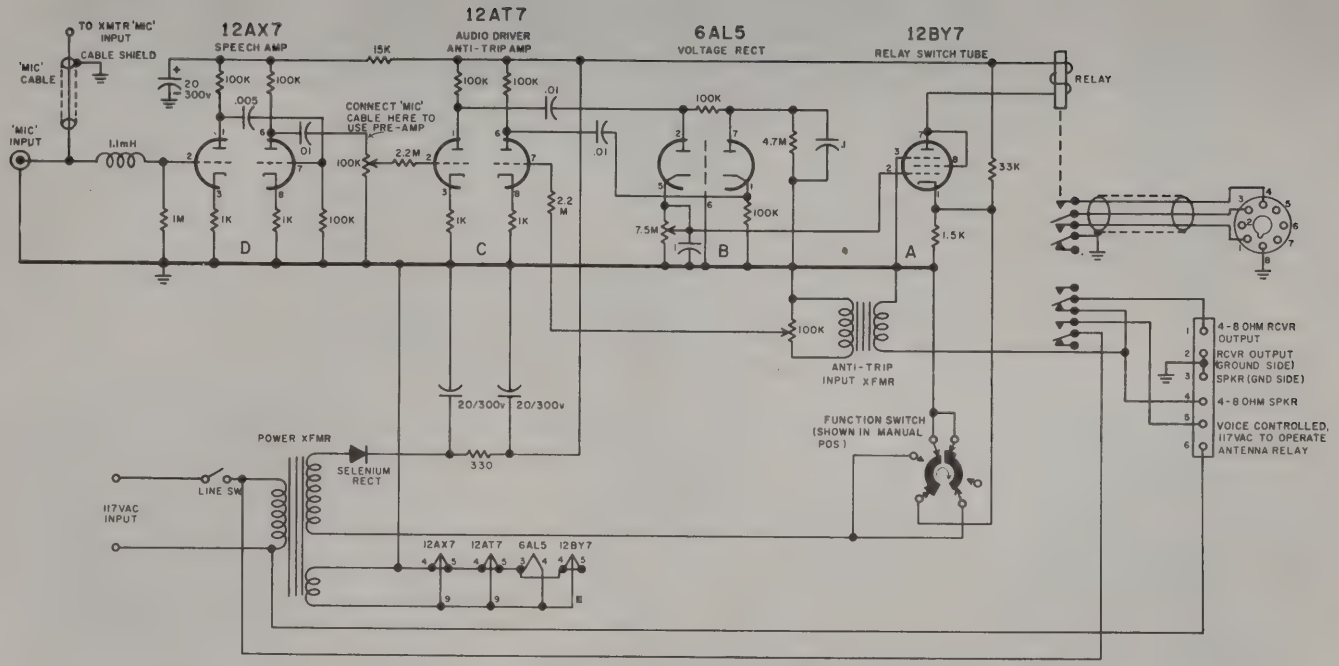


Fig. 8.4A—Schematic of the Heath VX-1, a typical voice operated relay. Optional connection allows the first stage (tube D) to be used as a microphone preamplifier, if desired.

The Globe Model VOX-10 contains an audio rectifier and relay control tube. The operating voltages are obtained from associated equipment. Provisions are made for adding optional anti-trip circuitry.

12AX7 tube. The same audio also goes to the microphone input on the transmitter. The speech amplifier is followed by a driver, using ½ of a 12AT7, which in turn feeds the diode audio rectifier, a 6AL5. Whenever the microphone is excited, a positive voltage will appear at pin 5 of the 6AL5. This voltage is directly coupled to the 12BY7 relay control tube to overcome the bias created by 33K and 1.5K resistors.

Audio from the speaker line connects to pin 1 of the terminal strip and passes through the VOX relay contact. A step-up matching transformer couples the audio to the second section of tube C, a 12AT7. The receiver audio drives pin 1 of the second section of the 6AL5, which is connected to produce a negative voltage. A 100K potentiometer between tubes C and D controls the amount of VOX voltage, while an identical potentiometer across the anti-trip transformer controls the amount of "opposition bias". A 7.5 megohm potentiometer in the pin 5 circuit of the 6AL5 determines how long the relay holds in after the speech ceases.

Although this unit is designed to work with the DX-100 and later Heath transmitters, it may be used with any AM or SSB unit.

8.5 PHONE PATCHES

Like the voice operated relay, the telephone patch is gaining wide popularity with the sideband operators. A phone patch is a device to connect the receiver output, and the transmitter input to the telephone lines. It most cases it is necessary to make a direct connection to these lines. The telephone company regulations specifically prohibit such connections, but most companies "overlook" such in-

All the operating controls are mounted on the front panel of the Heath VX-1 Electronic Voice Control unit. The time delay adjustment controls the length of relay hold-in.

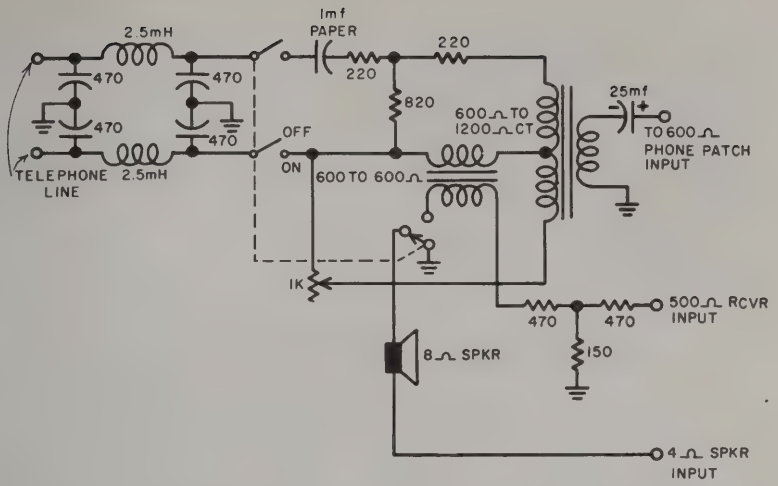


Fig. 8.5A—Hybrid type phone patch suitable for use with SSB equipment. This circuit is used in conjunction with the KWM-1.

fractions since they obtain revenue from long distance calls that would not otherwise be made. As long as the phone patch is "blocked" with a large *paper* capacitor, detection of the "illegal" connection is not likely.

One popular form of patch is known as the hybrid. A typical hybrid phone patch is shown in Fig. 8.5-A, and is used in the control box for the Collins KWM-1. The three section switch is used simply to turn the unit off-and-on. Note that the receiver output and the transmitter are connected to the hybrid transformer at the same time. Ordinarily the receiver audio would cause the VOX relay to go off like a machine gun. It would be far too much audio for the anti-trip circuit to overcome. The hybrid circuit overcomes this difficulty nicely. The receiver audio is connected between the center-tap of the transformer and one side of the phone line. Two audio paths, each with the same phase, go to the outside ends of the transformer primary. The 1K potentiometer assures that the same audio amplitude exists at each end of the transformer. Thus, with identical phase and amplitude at each end, perfect cancellation occurs in the secondary. In this manner, the receiver audio is balanced out of the transmitter input. An audio path from the receiver isolation transformer, to the phone lines does exist, however. Voltage induced in the secondary of the 1:1 transformer reaches the phone lines in series with the winding, the 220 ohm resistors, and the 1 mfd. paper capacitor. In exactly the reverse manner, audio from the phone lines is coupled through the hybrid transformer to the phone patch input.

The reason for these complications is to isolate the receiver output from the transmitter input. Thus, the operator can sit back and let the party on the other end of the wire trip the VOX circuitry. However, it has been the author's experience that most long distance phone lines are lower in level than local lines and the VOX does not trip properly. The ham usually winds up having the party say "over", and then he flips the switch manually.

For this reason, the author prefers the manually operated phone patch shown in fig. 8.5-B. The feedback loop between the receiver and transmitter is eliminated by manually disconnecting one, and connecting the other, as desired. The uncompli-

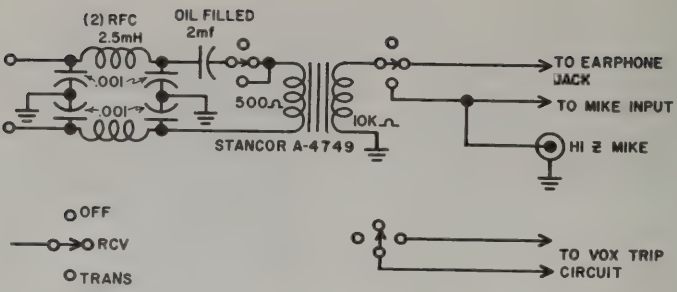


Fig. 8.5B—Manually operated phone patch suitable for use with SSB equipment. Unlabeled parts are the same as in fig. 8.5-A.

cated circuitry and the less expensive transformer make this an attractive construction project.

R-F filters and dc blocking capacitors are used in both circuits. The telephone lines pick-up a tremendous amount of r-f and rectification can occur in the patch. The purpose of the 1 mf blocking capacitor was mentioned earlier. An additional r-f by-pass should be connected directly across the carbon microphone button to prevent rectification at this point. If you do not care to tamper with the telephone company instrument, have them install the filter. A ceramic disc about .005 mf is quite satisfactory.

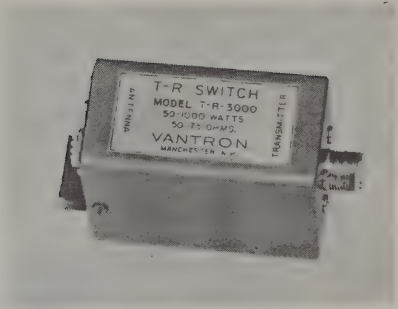
The hybrid transformer is quite expensive because the two halves of the primary must be perfectly matched in all respects to obtain a complete null. A push-pull 1200 ohm center-tapped transformer would *not* be satisfactory. Hybrid transformers are available in telephone company salvage or from most phone patch manufacturers. The C-161 loading coils described in the November 1957 issue of CQ Magazine (Hybrid Husbandry, page 52) should be quite satisfactory. The Stancor A-4749 used in the manual patch is available from Chicago Standard Transformer Corp. distributors.

8.6 ANTENNA (TR) SWITCHES

The purpose of the TR switch is to couple the receiver and transmitter to the antenna, electronically. In the usual installation, the antenna is connected directly to the link or output of a pi-network in the final. The receiver input is also connected to the antenna, but through an electronic switching circuit that "opens" when transmitting.

The switch has quite an impossible task to perform. It must reduce the transmitter output from several hundred volts, down to a low enough level to avoid ruining the receiver front end.

You must remember that the transmitter tank is always connected across the receiver input and acts as a tightly coupled absorption trap. Thus, at some frequency (usually the one you are working) the signals received by the antenna are effectively destroyed by the transmitter *suck-out*. This effect appears to be unbeatable. Since the suck-out can lower the received signal strength by as much as 20 db a marked reduction in receiver performance will often be noted. The amount of attenuation and suck-out frequency will vary from band-to-band. Switch manufacturers often neglect to mention this fact, and instead describe how much gain their switch has.



The Vantron TR switch is extremely compact and can be used to electronically disconnect the receiver input, when transmitting.

Often, the gain figures are for a switch that is not connected to the transmitter. Thus, a switch that is rated at 0 db gain on 10 meters might actually exhibit a -20 db gain in a completed installation. Additional gain in the switch will not help the situation either. The signal has already been removed from the system by the transmitter tank. Additional gain will bring up the signal and noise equally.

Before I incur the wrath of the switch manufacturers, it should be stated that it is possible to "juggle" the length of line between the TR switch and the final tank, in order to move the resonance (or suck-out) outside the band you are working. It may even be possible to eliminate the resonance on several bands, in some systems.

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Govier, **"10 Meter SSB with the BC 458,"** CQ, October, 1957, p. 48.

Arvonio, **"An Electronic Transmitter-Receiver Antenna Switch,"** QST, October, 1957, p. 32.

Bernard, **"Let's Increase VFO Stability,"** QST, October, 1957, p. 40.

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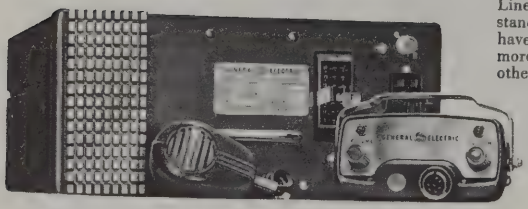
General Electric's Transistorized Progress Line offers highest performance . . . plus standby battery drain so low that you never have to turn off your TPL unit . . . fits in more places, in more positions, than any other two-way mobile radio you can buy.



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Since 1955 General Electric's tubed Progress Line has served customers with economical, high power, high reliability two-way radio. A wide range of models satisfies every mobile communication requirement. For field-proved performance and low maintenance cost even under the most rugged conditions, specify the Progress Line.

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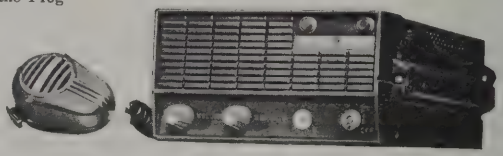

Many successful General Electric mobile radio authorized service stations were founded by licensed radio amateurs, and many now utilize the skills of hams on a part-time or full-time basis. Find out how you can become a General Electric service technician. Write National Service Manager, General Electric Company, Communication Products Department, Section 31, Mountain View Road, Lynchburg, Va.

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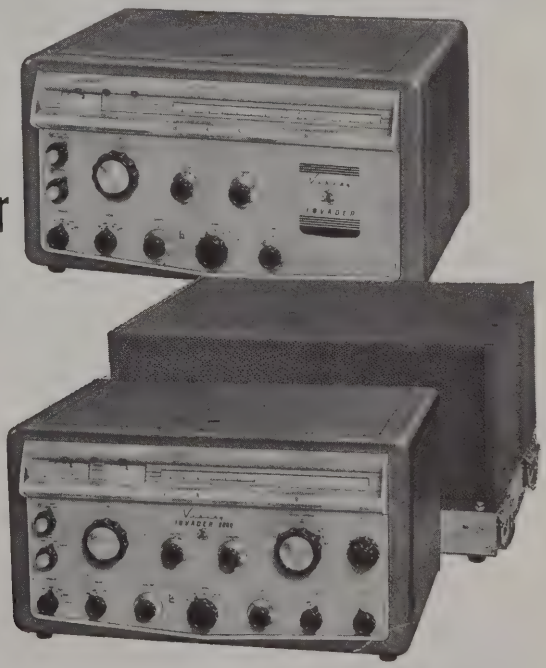

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Here's unsurpassed performance on all three . . . AM-CW and SSB! The Model 5100-B gives you:

- *high level push-to-talk AM telephony*
- *clean CW keying—break-in on all bands*
- *sparkling SSB combined with the 51SB-B companion sideband generator*

Check *all* the features built into this fine transmitter: input power of 180 watts CW-SSB, 140 watts AM phone; bandswitched throughout with integral VFO or crystal frequency control; coverage of 80 through 10 meter amateur bands; *plus* unitized construction, pi-network final, and integral low-pass filter. Skillful layout and imaginative circuitry assure effective transmission of signal with minimum harmonic content. Functional design within the r-f section, careful shielding, filtering, and by-passing are your assurance of the most dependable TVI suppression.

Trouble-free operation, ease of control and tuning, coupled with excellent quality of signal output with the Model 5100-B are sure to meet the exacting requirements of the most discriminating operator.

FOR OUTSTANDING SSB PERFORMANCE on the same amateur bands, combine the 5100-B with the completely bandswitched 51SB-B companion sideband generator. Powered by the transmitter, the 51SB-B can easily be hooked up to the 5100-B in less than a half hour. Tuning and operation are a breeze, with no test equipment required for installation or operation. And you'll begin immediately to enjoy such features as: voice operated control, push-to-talk, speaker deactivating circuit, true unitized construction, and TVI suppression.

*Net Price { Model 5100-B Transmitter.....$525.00
Model 51SB-B SSB Generator.....$265.00

FOR SPARKLING SSB WITH YOUR PRESENT TRANSMITTER . . . THE B&W MODEL 51SB SIDEBAND GENERATOR

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Here is a piece of commercial gear . . . to fascinate the discriminating amateur. Brilliantly designed—as modern as tomorrow—the REL L-103 1 KW Linear Amplifier sets new standards of performance and achieves the distinction of proven, measurable power output of 1000 watts PEP, CW, SSB, and FSK.

This completely self-contained, two-tube grounded grid linear amplifier measures only 7" high, 15" wide and 12" deep—yet gives you the big signal you want—consistently.

Matching power supply provides plate power and regulated screen voltage for the amplifier.

CHECK THESE FEATURES!

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- ✻ Built to MIL Specifications
- ✻ VSWR and Power Output Meter
- ✻ Peak Limiter Indicator
- ✻ Hi-Lo Antenna Output Switch
- ✻ High Efficiency Grounded Grid Circuit
- ✻ 4CX300A Beam Power Tube
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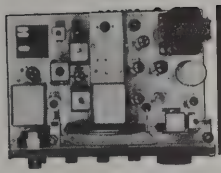
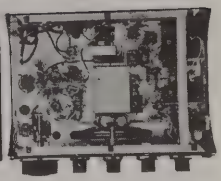


INSIDE THE

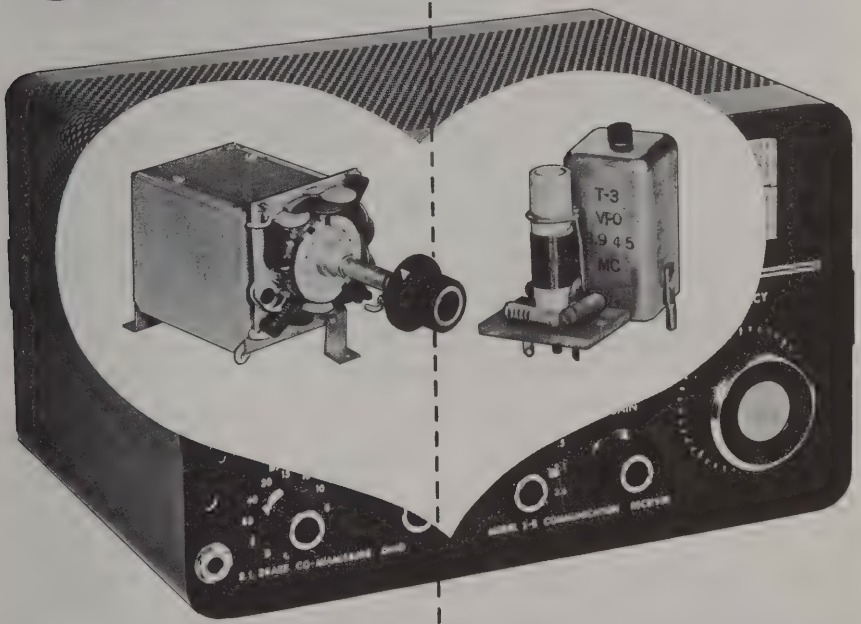
DRAKE 2-B RECEIVER



Essentials for SSB operation:

SELECTIVITY

STABILITY



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Newly designed, with .5 kc, 2.1 kc, and 3.6 kc switchable bandwidths. Continuous passband tuning functions on each bandwidth. All three bandwidths are steep-sided and can be used for SSB . . . the 3.6 for Hi-Fi SSB, the 2.1 for crowded band SSB and the .5 for extreme interference conditions.

VARIABLE FREQUENCY OSCILLATOR

This low frequency VFO, with latest materials and techniques for improved resistance to high humidity and temperature, maintains the 2-B's crystal-controlled frequency stability . . . less than 400 cycles warm up; less than 100 cycles after warm up; less than 100 cycles for plus or minus 10% line voltage change.

With the Model 2-B you have all the advantages of a "Ham Band Only" receiver. Superior SSB, AM, and CW in seven band switch positions . . . plus . . . five extra positions on the bandswitch for interchangeable plug-in crystals to permit reception of any 600 kc bands in the 3.5 to 30 mc range. Amateur Net, $279.95

Write for free brochure on the 2-B Receiver:

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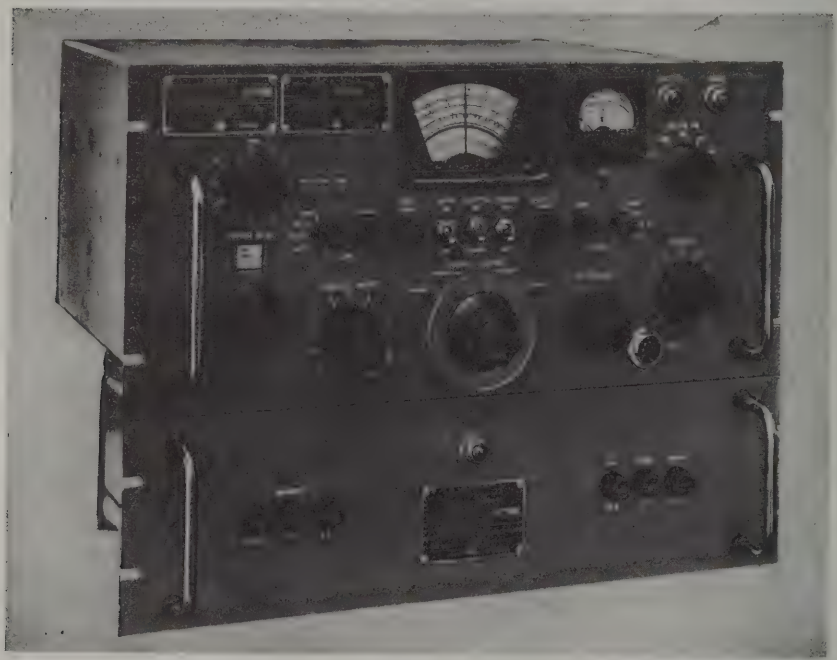
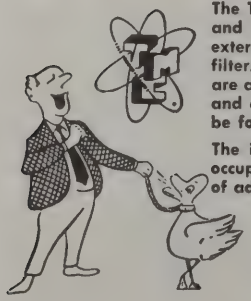


AN/URA-23A

MODEL SBE-2

AM • CW • SSB • DSB • ISB • FSK



The TMC SBE-2 is continuously tuneable in the range of 2 to 32 megacycles and is frequency controlled by means of temperature controlled crystals or external VFO. Sideband selection is accomplished by a specially designed filter. The carrier may be suppressed to 55 db. Harmonic and spurious output are at least 40 db down. VOX (voice control with anti-trip features) is built-in and adjustable. The SBE-2 provides at least 2.5 watts PEP output and can be followed by any appropriate AB-1 amplifier.

The illustration shows the basic exciter unit, with power supply. The unit occupies 8¾" of a standard WE relay rack. The power supply occupies 5¼" of additional space.

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The Completely New S/Line from Collins. The Latest addition to its distinguished single sideband series of amateur radio systems.

THE COLLINS 32S-1 TRANSMITTER $590.00

The 32S-1 is an SSB or CW transmitter with a nominal output of 100 watts for operation on all amateur bands between 3.5 and 29.7 mc. Input power is 175 watts PEP on SSB or 160 watts on CW.

Frequency Range: 80, 40, 20, 15, and 10 meter amateur bands. Easily retuned to frequencies between amateur bands by using different crystals.

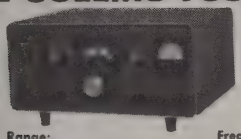
Output impedance: 50 ohms.

Frequency stability: After warm-up over-all stability due to temperature humidity, pressure and voltage variation is 100 cps.

Calibration accuracy: 1 kc.

Oscillators: Double conversion circuit is used with CR-18/U crystals in the HF oscillator. A VFO tuning 2.500 to 2.700 mc, provides 200 kc bands. A crystal oscillator operating on either side of the Mechanical Filter passband provides carrier for SSB generation and choice of upper or lower sideband.

THE COLLINS 75S-1 RECEIVER $495.00



The 75S-1 provides SSB, CW and AM reception on all amateur bands between 3.5 and 29.7 mc. It is capable of coverage of the entire HF spectrum between 3.5 and 20 mc by selection of the appropriate high frequency beating crystals.

Frequency Range:
80 meters—3.4 to 4.0 mc.
40 meters—7.0 to 7.4 mc.
20 meters—14.0 to 14.4 mc.
WWV—14.8 to 15.0 mc.
15 meters—21.0 to 21.6 mc.

Choice of three 200-kc portions of 10 meters: 28.5 to 28.7 furnished.

Overtravel—7.5 kc on all bands.

Frequency Stability: After warm-up, over-all stability due to temperature, humidity, pressure, and voltage variation: 100 cps.
Calibration accuracy: 1 kc.

Visual Dial Accuracy: 200 cps all bands.

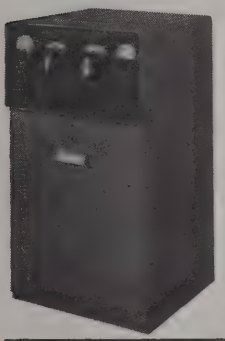
Electrical Dial Accuracy: (after calibration): 300 cps all bands.

Backlash: Less than 50 cps.

Sensitivity: The CW sensitivity is better than 1 microvolt (with a 50-ohm dummy antenna) for a 10 db single-plus-noise-to-noise-ratio.

Selectivity: 2.1 kc Mechanical Filter for SSB; 0.5 ks. Mechanical Filter (not supplied) for CW; 4.0 kc IF transformer passband for AM.

THE COLLINS 30S-1 LINEAR AMPLIFIER $1470.00



The 30S-1 Linear Amplifier rounds out the S/Line to make a single, complete, high powered amateur SSB station.

Frequency Ranges: 3.5—4.0 mc; 7.0—7.3; 14.0—14.4; 21.0—21.45; 28.0—29.7. Covers entire spectrum from 3.5 to 30 mc by retuning cathode circuit.
Output Impedance: 50 ohms.
Input Impedance: 50 ohms unbalanced.
Power Input: SSB-1 kw average, CW-1 kw.
Power Output: SSB: 1000 watts PEP with 40 db signal to distortion ratio; 1300 watts PEP with 35 db signal to distortion ratio.
CW: 600 watts with 1 kw input.
Controls: Band Change, Multimeter, Filament, H.V., Bias Control, Tuning, Loading.

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ACCESSORIES:

312B-4 SPEAKER CONSOLE integrates the 32S-1, 75S-1 and accessories into an operating system. $185.00

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- Extends effective usefulness of any receiver to 225 megacycles
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Range: 48.4 to 54.2 MC; 143.4 to 149.2 MC; 219.4 to 225.2 MC.
Noise Figure: 50 MC—2.5 db; 144 MC—4.0 db; 220 MC—6.0 db.
Calibration: Direct, MC subdivided in 100 KC divisions.
Panel Controls: Antenna changeover switch, band selector, tuning control, line switch.
Dimensions: 16½" wide, 10" deep, 10" high.
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The key to the superior performance of the 75-watt PL-177A, the 125-watt PL-4E27A, the 400-watt PL-175A, and the 1000-watt PL-172, is the suppressor grid. The longer plate voltage swing which can be obtained at a given screen voltage is the direct result of the electron beaming due to use of the suppressor, and leads to increased efficiency and output.

If you're buying a new transmitter or amplifier rather than building, look for Penta tubes in the equipment you buy.

Write for data sheets for any or all of these well designed, rugged Penta pentodes. Ask, also, for your free copy of "Transmitting Tubes for Linear Amplifier Service," which explains in detail why pentodes are your logical choice for single-sideband use.

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Type	FILAMENT Voltage (Volts)	FILAMENT Current (Amps)	Max. Plate Dissipation (Watts)	USEFUL OUTPUT* CLASS-AB1 LINEAR AMPLIFIER Plate voltage in volts 1000	1500	2000	2500	3000
PL-177A	6.0	3.3	75	96W	140W	210W	—	—
PL-4E27A	5.0	7.5	125	—	—	220W	280W	—
PL-175A	5.0	14.5	400	—	—	445W	570W	680W
PL-172	6.0	7.8	1000	—	—	1040W	1260W	1590W

*Actual power output delivered to load from typical amplifier.

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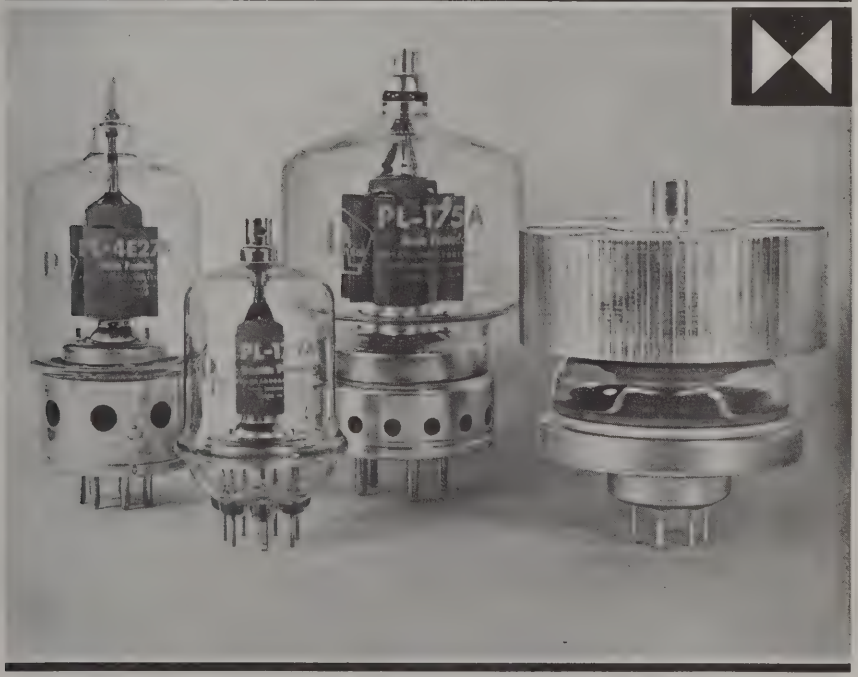

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THESE MULTIPHASE EXCITERS PIONEERED AMATEUR SSB

MODEL 20A

MODEL 10B — 10 watts PEP. Plug-in coils 160 thru 10 meters. Perfect voice control on SSB—DSB—AM and PM — CW breakin. Carrier and calibrate level controls. 40 DB suppression.

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MODEL 20A — 20 watts PEP. Bandswitched 160 thru 10 meters. SSB—DSB—AM—PM and CW. Magic eye monitors carrier null and peak modulation. Ideal for driving AB1, AB², and most Class B linears.

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Now MOBILE or FIXED STATION Electronic Antenna Change-Over Switches in one unit!

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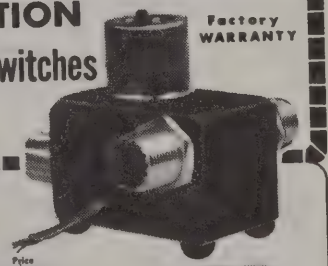

Now available for first time: DKC-TRM, 1.5 to 80 mc and DKC-TR2, 144 to 148 mc. Dow Key GUARANTEE, superb performance and quality. Exceptional ABOVE UNITY gain in both models.

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Only one ⅜" hole is needed, no screws to install this handy Dow-Key silver plated PANEL MOUNT UHF connector. Each 70c

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Models. 10.90 to 14.75

Positively Guaranteed to operate without AC hum or chatter. High-contact pressures possible with Dow-Key magnet principle, a new concept of low-contact resistance. Durable precision built for long life. Highest standard in performance and quality. Automatice receiver protecting connector (Patented). SPDT and DPDT external switches optional.

- See your local electronics dealer or write direct for complete Dow-Key specifications and catalogue sheet.

DOW-KEY CO. INC. Thief River Falls, Minn

Miller

At Your Local Jobber

R. F. COILS FOR ALL YOUR NEEDS

R.F. Chokes with Axial Leads

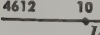
Dimensions: Phenolic form 7/16" diameter x 3/4" long. Iron core form 3/32" diameter x 7/8" long. Leads are 1½" long. Inductance tolerance: plus or minus 5%.

Phenolic Form
Single Layer Windings

Cat. No.	uh.	Ohms	ma	List Price
4580	0.1	.01	300	$ .55
4582	0.15	.012	300	.55
4584	0.22	.017	300	.55
4586	0.33	.019	300	.55
4588	0.47	.022	300	.55
4590	0.68	.03	300	.55
4592	0.75	.033	300	.55
4594	0.82	.035	300	.55
4602	1.	.05	300	.55
4604	1.5	.08	300	.55
4606	2.4	.16	300	.55
4608	3.9	.5	300	.55
4609	5.5	.69	300	.60
4610	6.2	.75	300	.60
4611	8.2	1.12	300	.60
4612	10	1.5	200	.60

Iron Core Form
Single Layer Windings

Cat. No.	uh.	Ohms	ma	List Price
4622	10	.06	300	$ .65
4624	15	.12	300	.65
4626	24	.28	300	.70
4628	39	.65	300	.70
4629	55	.92	300	.75
4630	62	1.0	300	.75
4631	82	1.6	300	.75
4632	100	2.0	200	.75



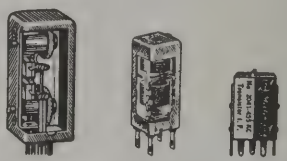
Phenolic Form
3-Section Windings

Cat. No.	mh.	Ohms	ma	List Price
4642	0.1	4.1	125	$ .75
4644	0.15	5	125	.75
4646	0.24	6.6	125	.75
4648	0.39	8.7	125	.80
4649	0.55	10	125	.80
4650	0.62	11	125	.80
4651	0.75	13	125	.80
4652	1.0	15	125	.80

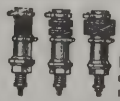
Iron Core Form
3-Section Windings

Cat. No.	mh.	Ohms	ma	List Price
4662	1	7	125	$ .85
4664	1.5	9	125	.90
4666	2.4	12	125	1.00
4668	3.9	17	125	1.10
4669	5.5	22	125	1.25
4670	6.2	33	100	1.25
4671	8.2	45	100	1.50
4672	10	47	50	1.75

Permeability Tuned I.F. Transformers



Cat. No.	Item	List Price
2031	455 KC Transister I.F.	3.00
2032	455 KC Transister I.F.	3.00
2041	455 KC Transister I.F.	3.00
2042	455 KC Transister I.F.	3.00
12-C1	455 KC Input I.F.	2.30
12-C2	455 KC Output I.F.	2.30
12-C9	455 KC Input I.F.	2.30
12-C10	455 KC Output I.F.	2.30
913-C1	455 KC Input I.F.	3.00
913-C4	455 KC Ouput I.F.	3.00
913-W1	1500 KC Input I.F.	3.00
913-W4	1500 KC Output I.F.	3.00
13-W1	1500 KC Input I.F.	2.50
13-W2	1500 KC Output I.F.	2.50
6203	4.5 MC Input Output I.F.	2.85
1463	10.7 MC Input Output I.F.	2.85
6190	21.25 MC Input Output I.F.	2.40



Adjustable Ceramic Forms and R.F. Coils

Permeability tuned coils wound on silicone impregnated ceramic forms. Hardware necessary for assembly is supplied.

Miniature Adjustable R.F. Coils

Dimensions (form): ¼" diameter x 7/8" long.

Cat No.	Microhenries	List Price
4500	Form only	$1.50
4501	.4- .8	2.00
4502	1.0- 1.6	2.00
4503	1.6- 2.8	2.10
4504	2.8- 5	2.20
4505	5 - 9	2.30
4506	9 - 16	2.40
4507	16 - 24	2.50
4508	24 - 35	2.50
4509	35 - 60	2.50
4511	60 -120	2.50
4512	110 -200	2.60
4513	190 -330	2.60
4514	320 -500	2.60
4514-1	480 -800	2.75

#4500 comes complete with hardware and core*

J. W. MILLER COMPANY
5917 South Main Street • Los Angeles 3, California

Canadian Representative: **Atlas Radio Corporation, Ltd. — 50 Wingold Ave., Toronto 10, Ont., Canada**

NEW GONSET SSB

GSB-100 TRANSMITTER

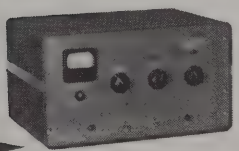
Introducing Gonset's big SSB value, the GSB-100. Completely self-contained with highly stable VFO and power supply, for operation on amateur 80, 40, 20, 15, 11 and 10 meter bands.

- Power input of 100 watts P.E.P.
- Operates on SSB with selectable sidebands.
- Operates also on AM, phase modulation or CW. Keying characteristics are excellent.
- Transmits both sidebands when on AM . . . avoids thereby, distortion present when carrier-and-one-sideband signals, at high modulation percentages, are received on conventional AM receiver.
- Frequency control is by fixed quartz crystal and exceptionally stable VFO *. Precise tuning is assured by dial assembly having gear ratio of 100:1.
- New Gonset FILTER-PHASING network gives excellent sideband rejection—quartz crystal band-elimination filter gives more than 60 db carrier suppression, avoids entirely all critical carrier balancing.
- Frequency coverage is full 600 kcs. within all amateur bands, 80 through 10 meters.
- Excellent voice operated control system (VOX). Biasing voltage is available for cut-off of external linear amplifier when receiving.

GSB-100 Transmitter . . . . . Model #3233

*11 meters and CW portion of 10 meter band covered by separate crystals not supplied.

Amateur Net $479.50



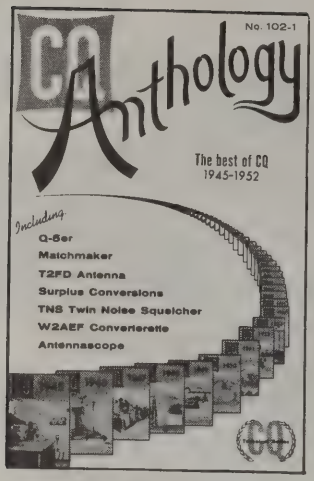
SSB LINEAR AMPLIFIER

New grounded grid linear amplifier is rated at 1000 watts peak input power. It is designed to operate with GSB-100, or similar SSB transmitters supplying 75–100 watts peak power drive. Amplifier is self-contained, includes power supply, pi-network output, antenna changeover relay. Provides bandswitched operation on 80, 40, 20, 15, 11 and 10 meters. Attractively styled. Same size cabinet and general appearance as GSB-100 Transmitter.

Model #3262

GONSET

GONSET DIVISION OF YOUNG SPRING & WIRE CORPORATION
801 SOUTH MAIN STREET, BURBANK, CALIF.



CQ ANTHOLOGY

We've looked back through the years 1945-1952 and assembled all in one place the articles that have made a lasting stir. The issues containing most of these articles have long ago been sold out and are unavailable. Th price is $2.00. PPD.

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HI-FI BOOK

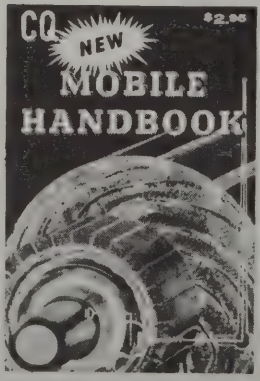
This nifty volume contains the latest dope on amplifiers, pre-amplifiers and equalizers plus a buyer's guide of component manufacturers!
Over 150—5½" x 8½" pages of heavily illustrated descriptions covering Hi Fi Audio Components — the greatest publication value in its field today. Only $2.50 PPD.

MOBILE HANDBOOK

This new Mobile Handbook by Bill Orr, W6SAI, has been getting raves from all of the experienced mobile operators. There is all sorts of information in here that cannot be found anywhere else. This is NOT a collection of reprints. $2.95 postpaid.

COMMAND SETS

This IS a collection of reprints, containing all of the available information on the conversion of the popular "Command" transmitters and receivers into good ham transmitters and receivers. Invaluable for Novice, Technician, General, Advanced and Extra class operators. 136 fabulous amazing terrific pages for only $1.50 PPD.



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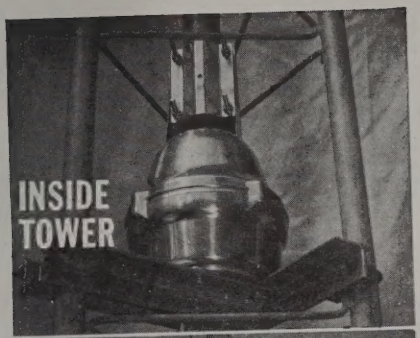
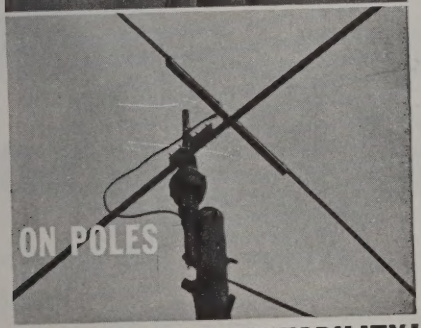


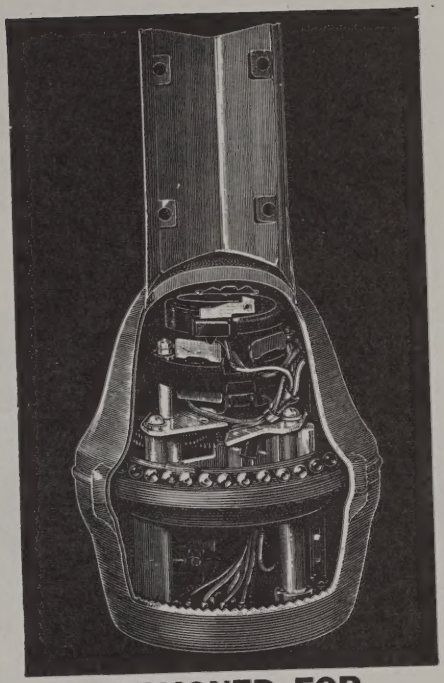


TALK ABOUT MOUNTABILITY!

Anyway you want to mount your antenna rotor, the HAM-M is the most versatile around! No special parts to buy! (Even the otherwise difficult inside tower mount can be accomplished with some angle iron and a hack saw) . . . Mountability like this is just one more reason why, at $119.50 amateur net, you just can't top the HAM-M! Ask your local CDE Radiart Distributor for all details.

CDE CORNELL-DUBILIER

CORNELL-DUBILIER ELECTRONICS, DIV. OF FEDERAL PACIFIC ELECTRIC CO., 50 PARIS ST., NEWARK 1, N. J.



DESIGNED FOR HALF-TON ANTENNAS

We've designed our HAM-M antenna rotors to support a dead weight of 1000 lbs. Your antenna probably weighs a small fraction of that, so see for yourself the kind of safety margin the HAM-M gives you!

But there's more! A positive electromechanical locking mechanism provides 3500 inch-pounds of resistance to the side thrust and whipping action of hurricane-force winds. And its bell-shaped, high tensile strength aluminum alloy housing is completely waterproof, assures brilliant performance even when caked with 5 inches of ice!

At $119.50 amateur net, the HAM-M is the greatest rotor value around! Ask your local CDE Radiart Distributor for all details.

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GLOBE ELECTRONICS

The Curtain is Going up on..

The All New

GLOBE ELECTRONICS CO.

AND IT'S NEW LINE OF AMATEUR RADIO EQUIPMENT

HG-303

Simple to tune. 3 Crystal positions. Compact and rugged. Functional. First quality in materials and operation. Reliability and durability guaranteed.

AMATEUR NET 99.95

HG-303M

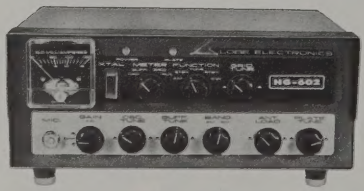
Styled and engineered to complement the HG 303 Transmitter. Allows conversion from CW to AM operation, comparable to finest quality transmitters.

PRICE TO BE ANNOUNCED

VFO-V10

The V 10 covers 6 through 160 meter bands. Complete with self-contained, well filtered power supply. Voltage regulated, temperature compensated for utmost stability. Easily read dial scales. R. F. output will Oscillator stage of any Transmitter on the market.

AMATEUR NET 69.95



HG-602

Functionally styled with complete simplicity of operation. Compact, entirely self-contained. Through TV I shielding. Instant band switching. Speech clipper/filter incorporated. 100% AM modulated. Silver-plated tank coils. 52 ohms link coupled output.

PRICE TO BE ANNOUNCED

Complete Specification Sheets Available
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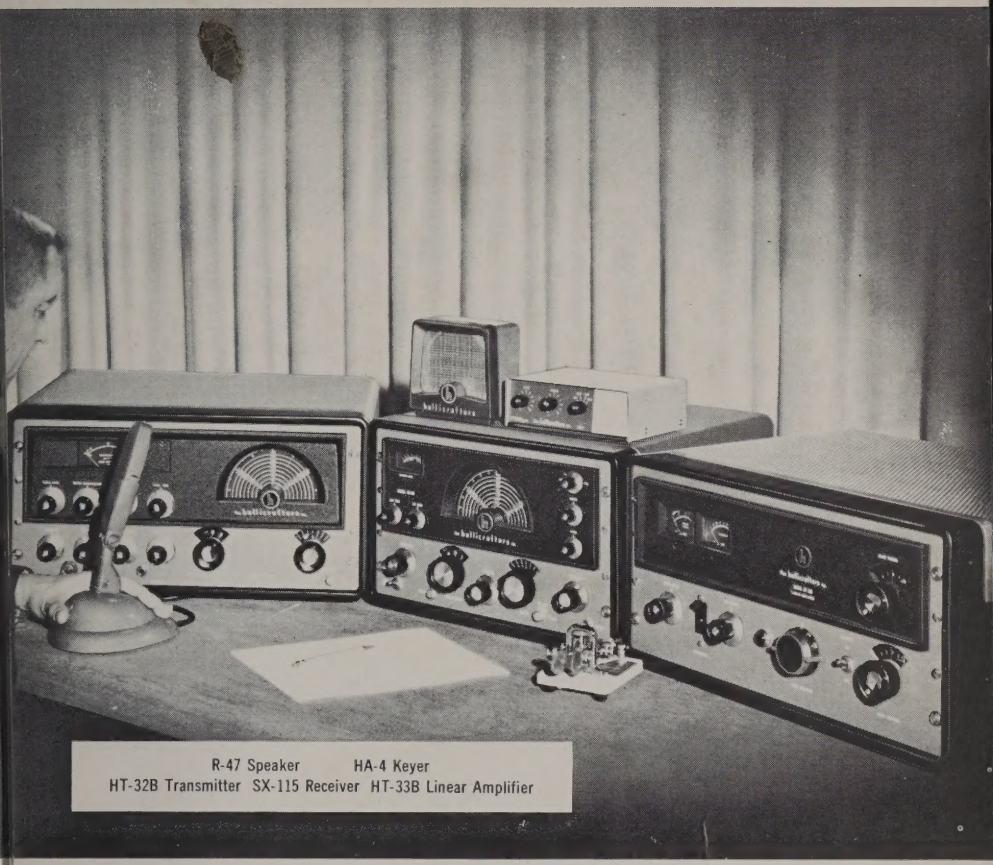

R-47 Speaker HA-4 Keyer
HT-32B Transmitter SX-115 Receiver HT-33B Linear Amplifier

The time-proven excellence of hallicrafters' HT-32B and HT-33B . . . the incomparable performance of the new SX-115 . . . the HA-2 and HA-6 transverters . . . and the fully transistorized HA-4 electronic keyer . . . now team up to bring you maximum flexibility and full coverage of 80 through 2 meters on SSB, CW and AM.

The new ideas in communications are born at . . .

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Chicago 24, Illinois

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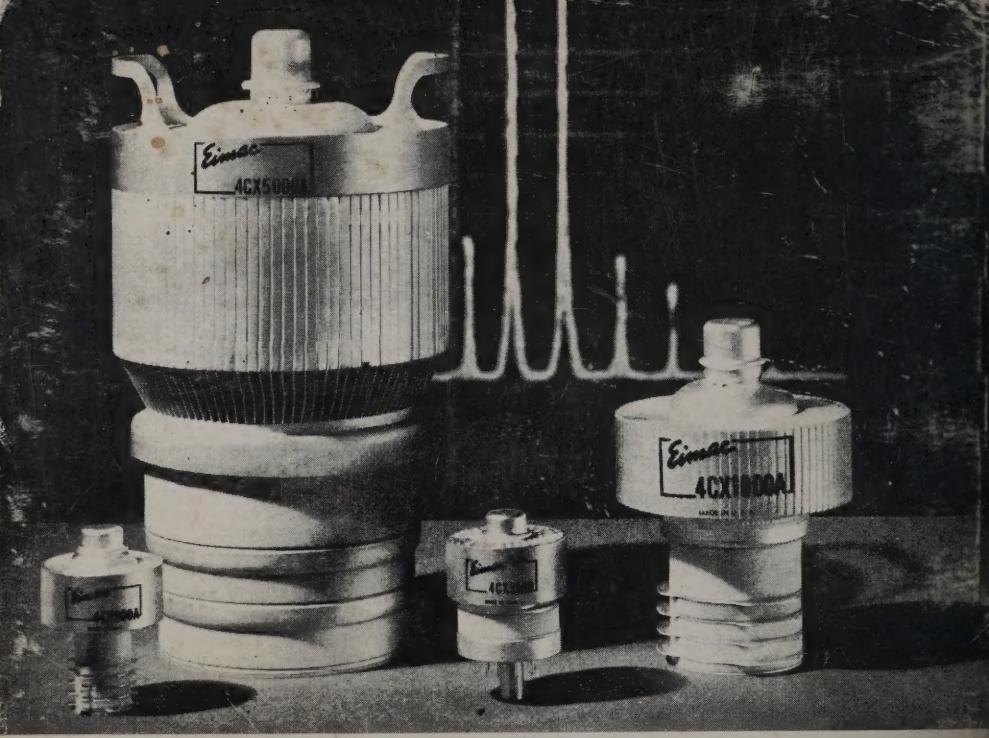

The Ideal Approach to SSB . . .

Eimac Ceramic Tetrodes for Class AB_1

Generating a clean SSB signal is one thing . . . amplifying it to the desired power level without distorting or broadening it is another. A modern class AB_1 final amplifier designed around an Eimac ceramic-metal tetrode is the ideal answer to the problem. Three of the four Eimac ceramic tubes shown above — the 4CX250B, the 4CX300A and the new 4CX1000A — are ideal for amateur radio application. All four offer the high power gain, low distortion and superior linear performance that is needed for class AB_1 operation. Each has performance-proved reserve ability to handle the high peak powers encountered in SSB operation. Efficient integral-finned anode cooler and Eimac Air System Sockets keep blower requirements at a minimum and allow compact equipment design. And all four incorporate the many advantages of Eimac ceramic-metal design, which assures compact, rugged, high performance tubes.

The reliability and performance that make Eimac ceramic tubes the choice for government and commercial transmitters can be yours in a compact, SSB tabletop kilowatt or a deluxe mobile rig. Eimac ceramic tetrodes simplify design and give more watt-hours per dollar.

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CLASS AB_1 SSB OPERATION

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Plate Voltage	2000 V	2500 v	3000 v	7500 v
Driving Power	0 w	0 w	0 w	0 w
Peak Envelope Power	325 w	400 w	1680 w	11,000 w